BERNARD BABANI

How To Use Op Amps

■ The operational amplifier is probably the most versatile IC available to the electronics engineer. For a price similar to a general purpose transistor, it is possible to purchase an IC with several hundred "components", very high gain and predictable performance. The Op Amp is thus a basic building block for applications from audio to industrial control.

■ This book has been written as a designer's guide covering many operational amplifiers, serving both as a source book of circuits and a reference book for design calculations. The approach has been made as non-mathematical as possible and it is hoped, easily understandable by most readers, be they engineers or hobbyists.

■ The text is divided into the following main chapters: Meet the Operational Amplifier, Basic Circuits, Oscillators, Audio Circuits, Filters, Miscellaneous Circuits, Common Op Amps, Power Supplies, Constructional Notes and Fault Finding.

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HOW TO USE OP AMPS

by

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Andrew Parr
April 1982

CHAPTER 1
MEET THE OPERATIONAL AMPLIFIER

1.1 INTRODUCTION

The Operational Amplifier is probably the most versatile circuit available to the electronics engineer. For a price similar to a general purpose transistor, it is possible to purchase an integrated circuit with several hundred "components", very high gain and predictable performance. The Op Amp is thus a basic building block for applications from audio to industrial control.

To many people the terms Op Amp and 741 are interchangeable, but in fact the 741 is merely the commonest member of a whole family of devices. This book has been written as a designers guide for most Operational Amplifiers, serving both as a source book of circuits and a reference book for design calculations. The approach has been made as non mathematical as possible, and should be understandable by most amateurs.

1.2 DC AMPLIFIERS

The designers of DC amplifiers using discrete components face a difficult task, and it is not difficult to see why. In Fig.1.1 we

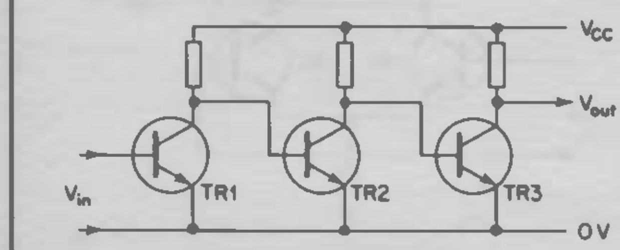


Fig. 1.1 Simple DC amplifier

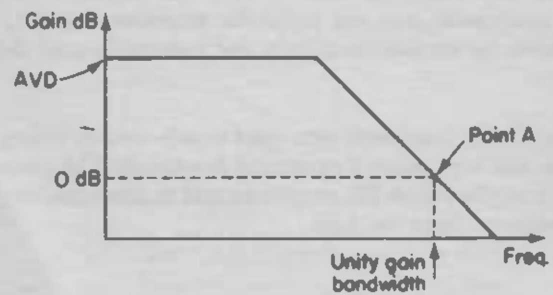


Fig. 1.5 Frequency response of DC amplifier

sufficient to drive the output to saturation, the output will change in a ramp form as Fig.1.6. The ramp slope is determined by the circuit of the amplifier, and is called the Slew Rate. A 741 has a slew rate of 1 volt/μs. High speed amplifiers have slew rates of the order of 100 volts/μs.

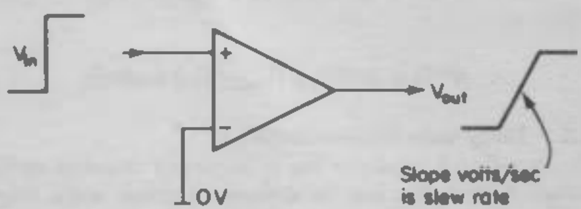


Fig. 1.6 Slew rate

1.3.4 Input Offset Voltage (V_{IO})

In Fig.1.7 we show all the relevant input and output currents. Suppose $V_1 = V_2 = 0$V. In theory V_{out} will be zero, but in practice due to slight manufacturing tolerances V_{out} will be

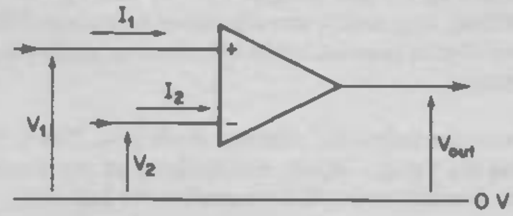


Fig. 1.7 Definition of offset voltages and currents

either positive or negative by a significant amount. If V_1 is slowly moved until V_{out} is zero, the difference between V_1 and V_2 is the input offset voltage. The figure for a 741 is around 2mV, quite small really.

1.3.5 Offset Voltage Temperature Coefficient ($\propto V_{IO}$)

In practice, V_{IO} is not particularly important since it can be nulled out by a zeroing potentiometer. What does usually matter is how V_{IO} changes with temperature. This is denoted by $\propto V_{IO}$ and is typically a few μV/°C.

1.3.6 Input Bias Current (I_{IB})

The transistors in the long tall pair need base current, hence in Fig.1.7, with $V_1 = V_2 = 0$V, I_1 and I_2 will not be zero. The bias current is defined as the average of I_1 and I_2 with V_1 and V_2 both zero. A typical value for a 741 is 0.1μA.

1.3.7 Input Offset Current (I_{IO})

With $V_1 = V_2 = 0$V in Fig.1.7, I_1 and I_2 will not be equal, and the offset current is simply the difference between I_1 and I_2. For a 741, the typical value is 20mA. Both I_{IB} and I_{IO} are temperature dependent, but for most circuits the effect is negligible. nA

1.3.8 Common Mode Rejection Ratio (CMRR)

A perfect DC amplifier amplifies the voltage between the + and – inputs, and totally ignores the common mode voltage on the terminals. The output voltages on Fig.1.7 with V_1 = 9 volts and V_2 = 9.001 volts, or V_1 = 0V and V_2 = 1mV should be identical. In practice, manufacturing tolerances will cause the amplifier to respond to the 9V common mode voltage to some extent.

First we must define the common mode gain. This is done by shorting the + and – inputs, and applying an input voltage as Fig.1.8. We then define the common mode gain as :–

$$ACM = \frac{\text{change in } V_{out}}{\text{change in } V_{in}}.$$

The change in signal is used to avoid problems with V_{IO} defined above.

We have already defined the DC gain, AVD, in section 1.3.1. The Common Mode Rejection Ratio is then simply:–

$$CMRR = \frac{AVD}{ACM}$$

CMRR is normally very large, so it is convenient to express it in decibels. A 741 has a CMRR of 90dB.

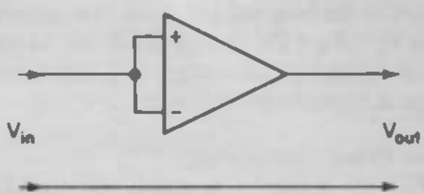


Fig. 1.8 Definition of common mode gain

1.4 STABILITY

An Op Amp has a frequency response that can, with some ICs,

give high gain at high frequencies. It is not particularly surprising to find many Op Amp circuits turn into Op Amp oscillators. There are many mathematical techniques to predict stability (Nyquist Diagrams, Bode plots to name but two) but these are beyond the scope of a book such as this. There are, fortunately, several empirical methods that can be used.

The first, and obvious, is not to use an amplifier that is too good for your application. Amplifiers such as the 741 have a strategically placed capacitor on the chip which causes the gain to roll off in a predictable manner. (The –3dB point on a 741 is as low as 10Hz, but the device still has useful gain at 40kHz, and unity gain at 1MHz). These devices are said to be unconditionally stable, and it is very unusual for a 741 and similar devices to misbehave.

If an unconditionally stable amplifier cannot be used, an Op Amp with compensation terminals should be substituted. These amplifiers have internal points within the circuit brought out so the user can shape the response. This can be done by two capacitors and a resistor as done on the 709 on Fig.1.9a or by a single capacitor as used by the 308 on Fig. 1.9b. Manufacturers data sheets give details of component values.

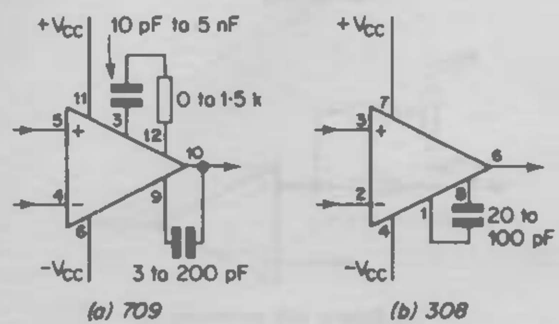


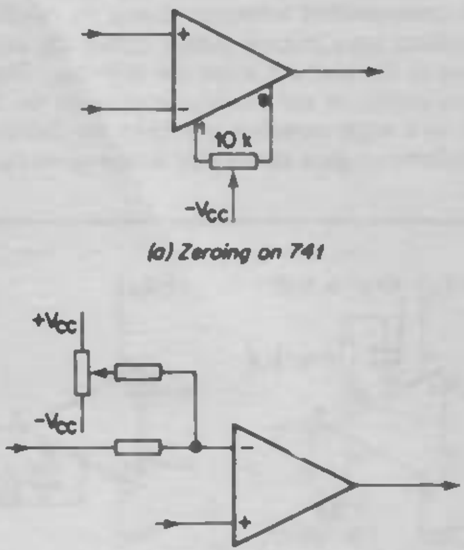
Fig. 1.9 External frequency compensation

1.5 ZEROING

In section 1.3.4 we saw that Op Amps have an input voltage offset of a few millivolts. In many low gain and AC applications this offset does not matter, but where a small voltage is to be amplified, the offset must be removed.

Many Op Amp ICs have internal zeroing points brought out to pins, typical being the 741 shown on Fig.1.10a. This allows the amplifier to be zeroed with one external potentiometer. As a cautionary note, shorting the zeroing pins to 0V or the positive supply is certain death for a 741.

An alternative method, useful for the inverting mode described in section 2, is to add a balancing voltage by a potentiometer and resistor as Fig.1.10b. This method should be used where a precision amplifier is required even if zeroing pins are provided, because the resulting zero is more stable.



(a) Zeroing on 741

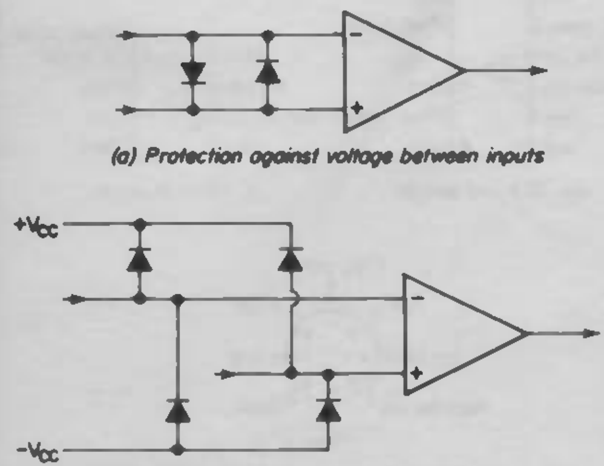
(b) Zeroing with balancing pot

Fig. 1.10 Zeroing of DC amplifiers

1.6 PROTECTION

Integrated circuit Op Amps are very well protected, and later versions are almost bullet proof. The 741 in particular seems immune to short circuits and will even stand reversed supplies and being plugged into a socket wrong way round!

Some Op Amps can be destroyed by a large voltage (above 5 volts) between their inputs. These amplifiers should be protected by diodes across the inputs as Fig.1.11a. In normal use the diodes do not conduct, and will not affect the circuit.



(a) Protection against voltage between inputs

(b) Protection against inputs going outside supplies

Fig. 1.11 Protection of DC amplifiers

Most Op Amps will be destroyed if an input goes outside the supply rails. If this is possible (remember Murphy's Law!) diodes should be added as Fig.1.11b.

FET Op Amps are vulnerable in the same way as CMOS, and normal precautions should be taken against static damage when using them.

1.7 PACKAGING

There is a remarkable degree of consistency in the packaging of Op Amps. The first Op Amp was the 709 shown on Fig. 1.12a. This configuration was carried forward to the 8 pin DIL packaging of the 741 shown on Fig.1.12b. It will be noted

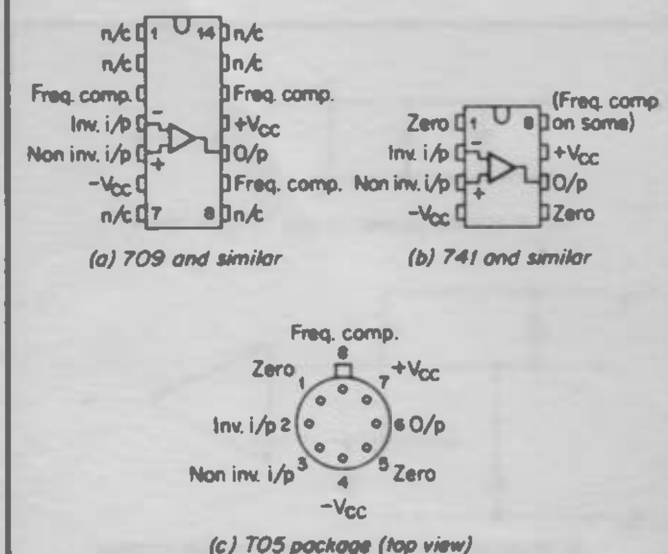


Fig. 1.12 Packaging

that this is compatible with the 709, and has, in fact, been adopted as a standard.

Wherever possible, manufacturers have used 1.12b, with minor variations for individual functions (such as zeroing and compensation) on pins 1 and 8. Almost all Op Amps use the same pins for inputs, supplies and output. This simplifies designs considerably as a better specification (or cheaper) Op Amp can easily be substituted if the initial choice is wrong.

Early Op Amps used the 8 pin TO5 package of Fig.1.12c. This has been largely superceded by the DIL package, and should only be used on a replacement basis.

1.8 COMPARATORS

A close relative of the Op Amp is the comparator. This is a

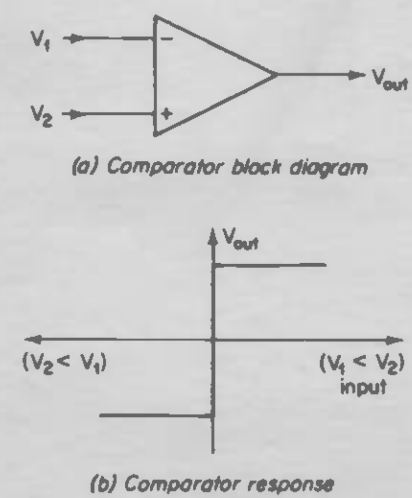


Fig. 1.13 Comparators

high gain DC amplifier designed to compare the magnitude of its two inputs. It thus has the response of Fig.1.13b. Comparators are, however, designed for speed, and most will switch in around 100nS.

Comparators have offsets similar to Op Amps, and in general the terms of section 1.3 apply to comparators. A typical comparator is the LM311.

CHAPTER 2.
BASIC CIRCUITS

2.1 INTRODUCTION

This book describes many practical circuits using operational amplifiers. In this section the basic building block circuits are described, and the design principles are given. Although described for the 741, they can of course, be implemented with any operational amplifier.

2.2 INVERTING AMPLIFIER

2.2.1 Description

The classic operational amplifier circuit is the inverting amplifier shown on Fig.2.1. Since the amplifier has a very high gain, and the output is going to stay within the supply rails we can say that both of the input pins (2 and 3) are going to stay within a millivolt of 0V. For all practical purposes, therefore, we can say that the junction of R1 and R2 is at 0V. This is known as a Virtual Earth in the jargon, and simplifies the design considerably.

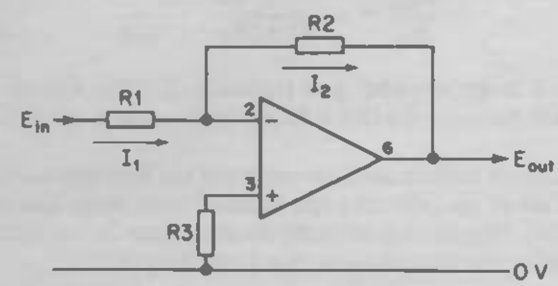


Fig. 2.1 Practical inverting amplifier

We can now work out I_1 easily, because:–

$$I_1 = \frac{E_{in}}{R1}.$$

Similarly, I_2 is given by:–

$$I_2 = \frac{E_{out}}{R2}.$$

Since $I_2 = -I_1$ we can write:–

$$\frac{E_{in}}{R1} = -\frac{E_{out}}{R2},$$

or

$$\frac{E_{out}}{E_{in}} = -\frac{R2}{R1}.$$

The minus sign denotes inversion (i.e. a positive input voltage gives a negative output voltage).

This is an interesting result, since the amplifier gain does not appear in the formula, and the gain is determined solely by R1 and R2!. This is, in fact, correct if the amplifier gain is very high and the closed loop gain relatively low. The full formula is:–

$$\frac{E_{out}}{E_{in}} = -\frac{R2}{R1 + (R2 + R1)/A}$$

where A is the amplifier gain (typically 20,000). For all practical purposes the (R2 + R1)/A term can be neglected.

In the above analysis we have neglected the base bias currents. These are of the order of a few hundred nano Amps and cause an offset. The effect is normally small and can be minimised by making the impedance at pins 2 and 3 equal.

i.e.

$$\frac{1}{R3} = \frac{R1 + R2}{R1R2}$$

Any remaining offset due to the amplifier offset voltage or input current offset can be trimmed out with a zeroing potentiometer if required.

The input impedance of the closed loop amplifier is simply the value of R1.

2.2.2 Designing an Inverting Amplifier

If a relatively low gain amplifier is required, with an input voltage above about 100mV the design procedure is straightforward. Choose R1, R2 and R3 from convenient values in the come-in-handy box. As a rough rule of thumb choose R1 to be about 10k and work on from there.

For lower input voltages and higher gains, slightly more care is necessary, and the following check list should be followed.

1. V_{IO} is not particularly important since it can be nulled out, but the change in V_{IO} with temperature is often critical. Work out $\propto V_{IO}$ over the expected temperature range (when in doubt use 30°C). This should be smaller than your input signal by a factor of at least ten. If not choose a better amplifier.
2. Check that the offset due to the bias current (I_{IB}.R1) is less than the input signal. If not, reduce R1 or choose a better amplifier. Any small offset can be trimmed out by the zeroing potentiometer.
3. Given R1, R2 can now be calculated using the formula given. Check now for temperature variations on the offset current $\propto I_{IO}$. The equivalent offset is $\propto I_{IO}$.R2. If this is unacceptably large choose a better amplifier or reduce R1 and start again.
4. Calculate R3 as described earlier.

The above analysis assumes that R1 is fed from a low impedance source. If this is not true, R1 should be replaced by ($R1 + R_S$) in the equations where R_S is the source resistance, or an interposing high input impedance buffer used.

2.3 NON-INVERTING AMPLIFIER

2.3.1 Description

The classic non-inverting amplifier is shown on Fig.2.2. By similar arguments to those in section 2.2 we can say that the voltages on pins 2 and 3 will be equal. By simple analysis, the voltage on pin 2 is given by:–

$$E_1 = E_{out} \cdot \frac{R1}{R1 + R2}$$

since $E_{in} = E_1$ as described above

$$E_{in} = E_{out} \cdot \frac{R1}{R1 + R2}$$

or

$$\frac{E_{out}}{E_{in}} = \frac{R1 + R2}{R1}.$$

To minimise the offset due to the bias current, R3 should be chosen such that

$$R3 = \frac{R1 \cdot R2}{R1 + R2}$$

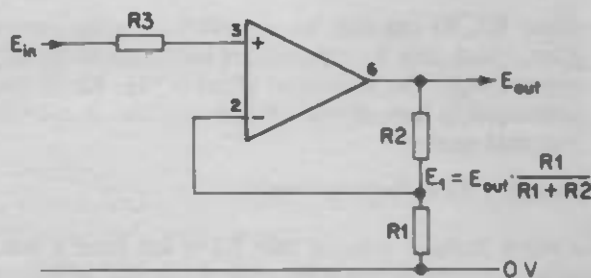


Fig. 2.2 Practical non inverting amplifier

The input impedance of the non inverting amplifier is very high.

2.3.2 Designing a Non-inverting Amplifier

Procedures 1 to 4 in section 2.2.2 apply to both inverting and non-inverting amplifiers.

2.4 BUFFER AMPLIFIER

A unity gain version of the non-inverting amplifier is shown in Fig.2.3. This acts as a "super" emitter follower, and has a very high input impedance. It is a very useful circuit where a buffer stage is required. In most applications the only design limitation is $\propto V_{IO}$.

The buffer amplifier is very effective when used with FET input amplifiers.

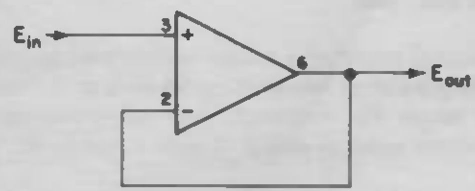


Fig. 2.3 Unity gain buffer

2.5 DIFFERENTIAL AMPLIFIER

The differential amplifier is used where it is desired to amplify the difference between two voltages, as summarised on Fig. 2.4. The output voltage is given by:–

$$E_{out} = -\frac{R_b}{R_a}(V_1 - V_2)$$

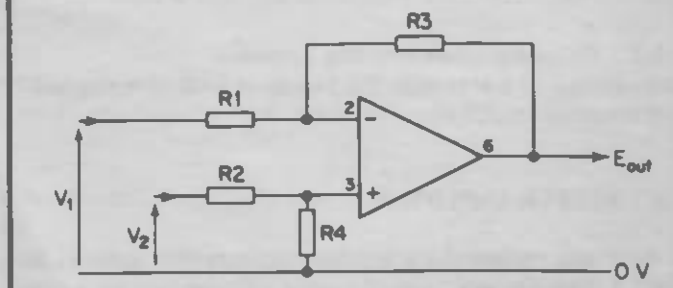


Fig. 2.4 Differential amplifier

if $R1 = R2 = R_a$ and $R3 = R4 = R_b$

Note that both input resistors have the same value, and R3 and R4 have the same value. This is essential if the circuit is to work properly. If all four resistors have the same value, the circuit has unity gain.

The differential amplifier is widely used to remove common mode noise present on low level signals such as thermocouples and strain gauges. For maximum rejections of common mode noise, precision resistors should be used to ensure R1 = R2 and R3 = R4.

The impedance seen by V_2 is simply R2 + R4. The impedance seen by V_1 is variable dependent on the signal level, but is of the order of R1. It follows that the source impedances of V_1 and V_2 must be significantly lower than R1 and R2.

2.6 ANALOG COMPUTERS

2.6.1 Introduction

The history of the analog computer is, perhaps, one of the great ironies of the impact of integrated circuits. In the 1950's, computers split into two distinct types. The first was the

digital computer, using binary logic gates to perform computational tasks. The second was the analog computer which used operational amplifiers.

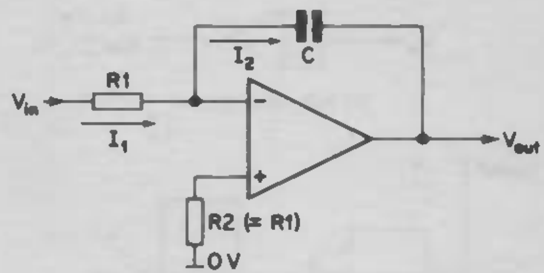
A digital computer's basic instructions are very limited (see the author's book "A Microprocessor Primer", published by Bernard Babani (publishing) Ltd, Book No.BP72) and these early computers were very slow. As a result they had great difficulty with problems involving the calculus operations of integration and differentiation.

An analog computer represents variables by voltages, and these can be continuously variable. It is very easy to perform calculus operations with operational amplifiers, so a model of the system under study is built. If we were building a model of a chemical plant, say, we would derive mathematical relationships for all the variables (flows, pressures, temperatures etc.). We could, for example, represent a temperature of 100 – 500°C by a voltage in the range 1 – 5 volts at an amplifier output. When completed the model could be studied and used as a test bed for instrumentation etc.

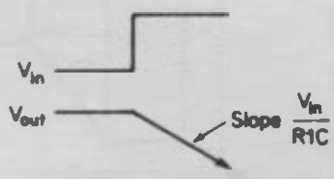
Analog computers were very popular in the 1950's and 1960's, but the technology of the time was very restrictive. Valve operational amplifiers drifted in an alarming manner and transistor versions required very careful design. Nonetheless most engineering design offices and universities had an analog machine tucked away somewhere, tended by a team of engineers because these machines were BIG as well as temperamental.

The arrival of integrated circuits should have revolutionised analog computers as cheap, small, and above all stable, DC amplifiers became available. Unfortunately the same technology changed digital computers by increasing their speed to the point where complex mathematical calculations could be done in a reasonable time.

Analog computers have now almost followed the "Dodo" into the history books, which is very sad because building an analog model of a process is very instructive. The technique of analog



(a) Integrator circuit



(b) Circuit operation

Fig. 2.8 Theoretical integrator circuit

A practical integrator is shown on Fig.2.9. With the values shown the output changes at 1 volt/second for a 1 volt step. The zeroing control RV1 is, in fact, more than a zero control since it is used to balance out the input bias current which would be integrated if left uncorrected.

RV1 should therefore be adjusted until V_{out} stabilises for zero volts in. In practice this will be impossible to achieve, and RV1 should be set for minimum drift. SW1 is therefore included to discharge C1 prior to use.

2.6.4 Differentiator

Differentiation is the calculus operation used to determine the rate of change of a varying signal. This is similar to working out the acceleration of a car from its observation of the speed.

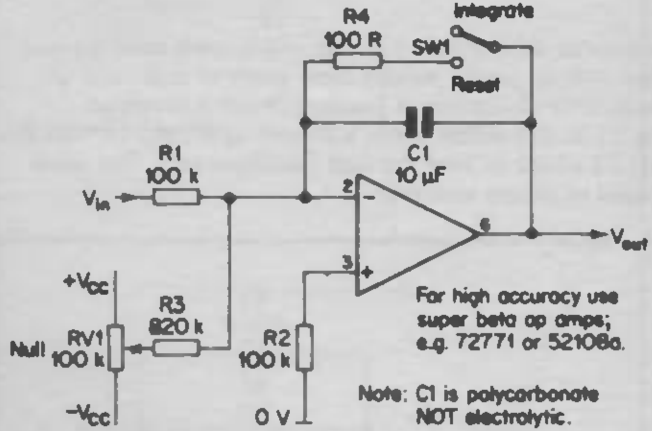


Fig. 2.9 Practical integrator circuit

The circuit for a differentiator is shown on Fig.2.10. By calculation similar to that in the preceeding section we find that.–

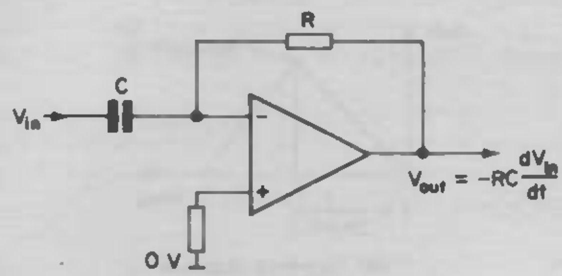
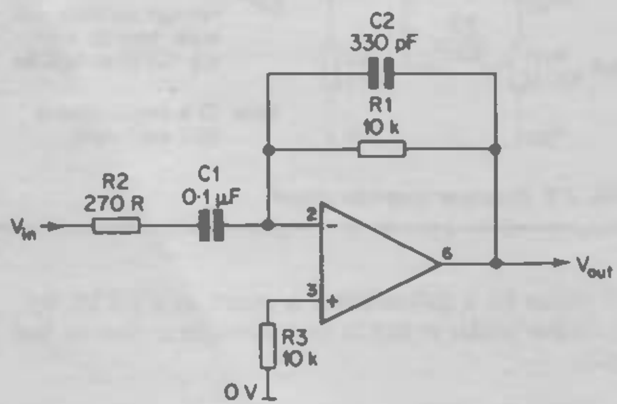


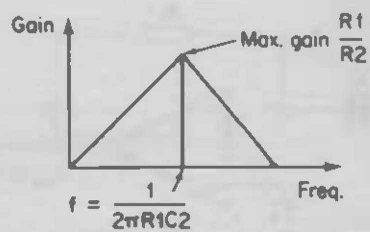
Fig. 2.10 Differentiator circuit

$$V_{out} = -RC\frac{dV_{in}}{dt}.$$

In practice differentiators are not widely used, since the gain rises with frequency making them prone to noise and unpredictable oscillations. A practical circuit is shown on Fig.2.11a. The differentiator is formed by R1 and C1 with R2 and C2 added to limit the high frequency gain. The values should be chosen such that:–



(a) Practical circuit



(b) Frequency response

Fig. 2.11 Practical differentiator

$$R1\ C2 = R2\ C1$$

The response is shown on Fig.2.11b, and exhibits the characteristics of a differentiator up to a frequency given by:–

$$f = \frac{1}{2\pi R1C2}$$

after which the gain falls. The maximum gain is R1/R2.

2.6.5 Multipliers and Dividers

One of the reasons for the demise of the analog computer was the difficulty of performing accurate multiplication and division. The circuits are complex, even with IC Op Amps, so only the principles will be given.

The simplest multiplier uses a pulse width and pulse height modulator, and is shown in block diagram form on Fig.2.12. The circuit starts with a sawtooth oscillator which is fed to a comparator along with the first input voltage X. The comparator output is thus a constant frequency square wave whose pulse width is proportional to X, and these are used to turn on transistor TR1.

The collector of TR1 is connected to the output of the buffer amplifier hence the collector switches between 0V and the second voltage Y. The resulting signal is inverted to give a constant frequency pulse train whose pulse width is proportional to X, and height to Y. A simple filter gives an output voltage proportional to X.Y.

A second, and faster, technique uses the fact that the relationship between the current through a diode and the voltage drop across it is given by:–

$$I = Ae^{BV} \text{ where A and B are constants.}$$

By use of V to I and I to V amplifiers and diodes it is possible to make logarithmic and antilog amplifiers. With the input voltages converted to logarithmic equivalents, multiplication

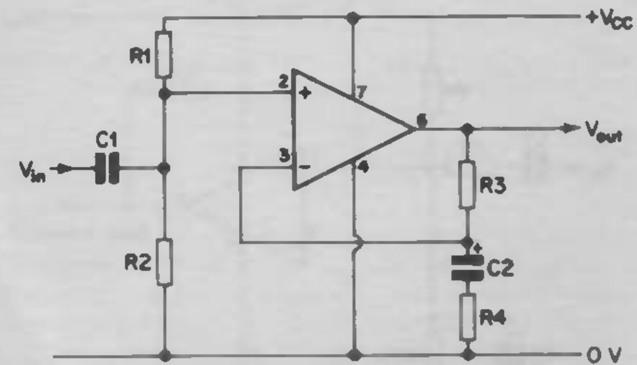


Fig. 2.16 Non inverting A.C. amplifier

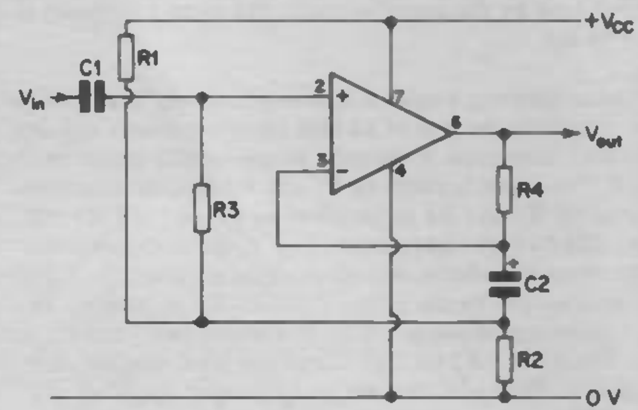


Fig. 2.17 Bootstrapped A.C. amplifier

inputs are equal in both amplitude and phase, the input impedance is very high. To a first approximation the input impedance is given by:–

$$R_{in} = \frac{\text{Amplifier Open Loop Gain}}{\text{Closed Loop Gain}} \times R3$$

All the above amplifiers have unity gain to DC offset effects, so there is no need to provide zeroing facilities.

2.8 SCHMITT TRIGGER

The Schmitt Trigger is a circuit widely used to convert a slowly varying signal into the crisp on/off signals used in logic and other digital systems. The circuit exhibits hysterisis, in that a considerable backlash is designed in. This is best summarised by Fig.2.18. The output thus is at either the positive supply or the negative supply and switches at the upper and lower trigger points. The backlash gives considerable protection against jitter on the digital output.

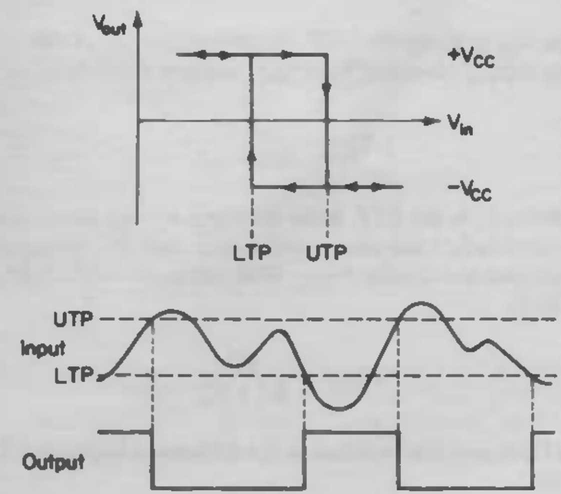


Fig. 2.18 Hysteresis and the Schmitt trigger

Schmitt triggers are available in ICs (e.g. 7414), but these have fixed trigger points. A versatile Schmitt trigger with adjustable trigger points can be made with an Op Amp and two resistors as shown on Fig.2.19.

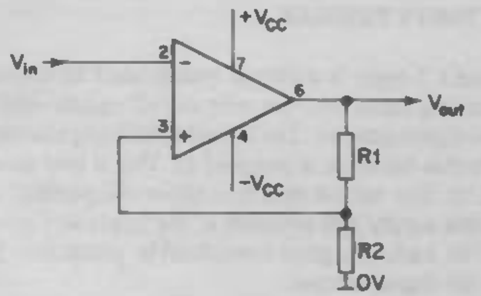


Fig. 2.19 Basic symmetrical Schmitt trigger

Suppose V_{in} is above the UTP; the output will be at the negative supply (denoted by $-V_{CC}$) and pin 3 will be at:–

$$-V_{CC} \cdot \frac{R2}{R1 + R2}$$

This obviously is the LTP, since the input voltage has to drop to this level before the circuit will switch, and the output goes up to the positive supply $+V_{CC}$. With the input below LTP, pin 3 will be at:

$$+V_{CC} \cdot \frac{R2}{R1 + R2}$$

This is UTP, and the backlash is the difference between LTP and UTP.

With equal positive and negative supplies (as will usually be the case) the trigger points will be symmetrical about 0V, and

the hysterisis will be twice the trigger points. If a non symmetrical response is required this can easily be provided by the circuit of Fig.2.20 where R2 is returned to a voltage of V_2 provided by the potentiometer RV1. It is important that the resistance of RV1 is lower than R1 + R2 by a factor of ten to ensure V_2 is stable.

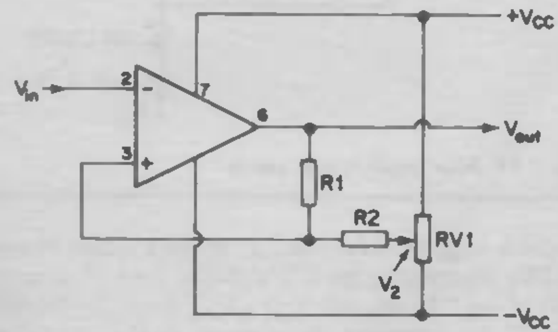


Fig. 2.20 Assymetrical Schmitt trigger

The mathematics of this circuit is slightly more complex, but the UTP is given by.–

$$\frac{R1V_2 + R2\,(+V_{CC})}{R1 + R2}$$

and the LTP by.–

$$\frac{R1V_2 - R2\,(-V_{CC})}{R1 + R2}.$$

Care should be taken in applying the correct polarities in the above equation.

If the supplies are not particularly stable, the trigger points may vary unacceptably. Zener diodes should then be used to

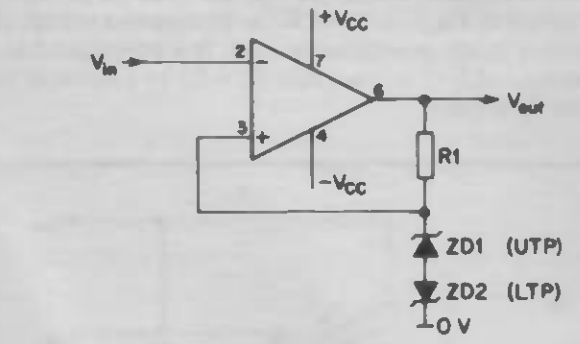


Fig. 2.21 Zener diode trigger points

give stable trigger points. Fig.2.21 shows a simple circuit, with ZD1 determining the UTP and ZD2 the LTP directly. The circuit of Fig.2.21 can only have a UTP above 0V and an LTP below 0V. An alternative circuit shown on Fig.2.22 can provide UTP and LTP of the same polarity. The operation of the

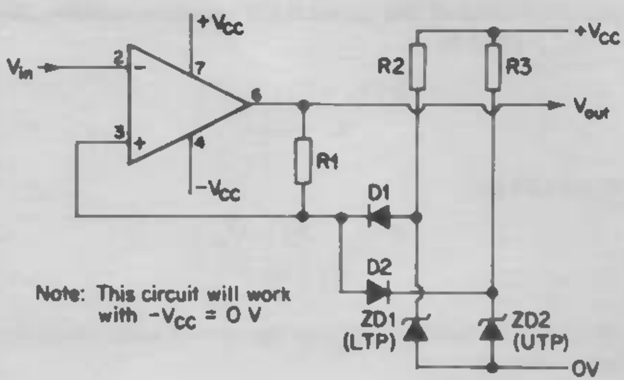


Fig. 2.22 UTP and LTP same polarity

circuit is straightforward, but the Zener bias resistors R2, R3 should be considerably lower in value than R1. This is no great problem, since R1 has to supply minimal current to the non-inverting input. Obviously ZD2 determines the UTP, and ZD1 the LTP.

It should be noted that some amplifiers (notably the 709) suffer from latch up when driven into saturation. Such amplifiers cannot be used in Schmitt trigger circuits.

In section 3.3 (Function Generator) a non-inverting Schmitt trigger is described as part of an oscillator.

2.9 INCREASED CURRENT OUTPUT

Most operational amplifiers can source, or sink current up to 10mA. This is quite adequate for many applications, but quite often significantly higher currents are required. Higher current outputs can be obtained by the use of additional components.

The simplest circuit uses an NPN and a PNP transistor connected as two emitter followers as shown on Fig.2.23. For higher currents Darlington connected transistors can be used,

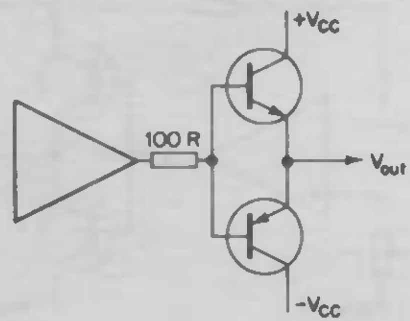


Fig. 2.23 Increased current output

either by means of four discrete transistors or with two Darlington ICs such as the TIP121 (NPN) or TIP126 (PNP). It should be noted that the use of booster transistors will reduce the available output range by about 1 volt in each direction (about 1.5 volts for Darlingtons).

If booster transistors are used, it is very important to take the feedback from the junction of the two emitters as shown on Fig.2.24. This will ensure correct circuit operation. Taking the feedback from the Op Amp output results in some very odd effects!

Variations on the circuit of Fig.2.23 will be found later in the Audio section, where they make useful and simple audio amplifiers.

The choice of transistors is determined mainly by the current gain. The Op Amp can source or sink 10mA, hence if the transistors in Fig.2.23 have a β of 20 (a reasonable figure for a power transistor) we could drive a 200mA load. With Darlingtons a β of 1000 is easily obtained, so we could, in theory, drive a 10 Amp load, although this would be

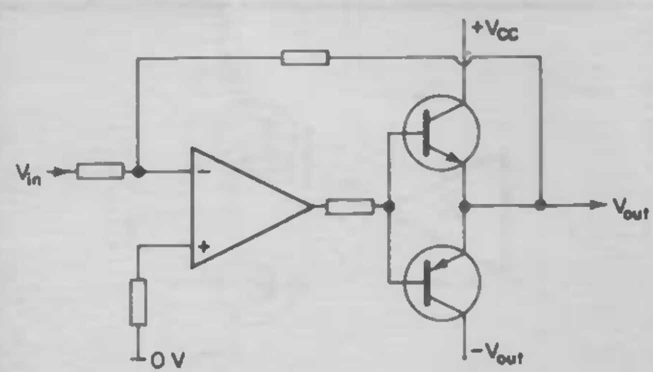


Fig. 2.24 Connection of feedback

impractical for other reasons! The output transistors will dissipate a fair amount of heat, the maximum dissipation occurring when the output is halfway between 0V and a supply rail. For example, if we are driving 500mA from a ±15V supply, the maximum dissipation in the output transistors will occur when the output voltage is 7.5V. The current will be 250mA,

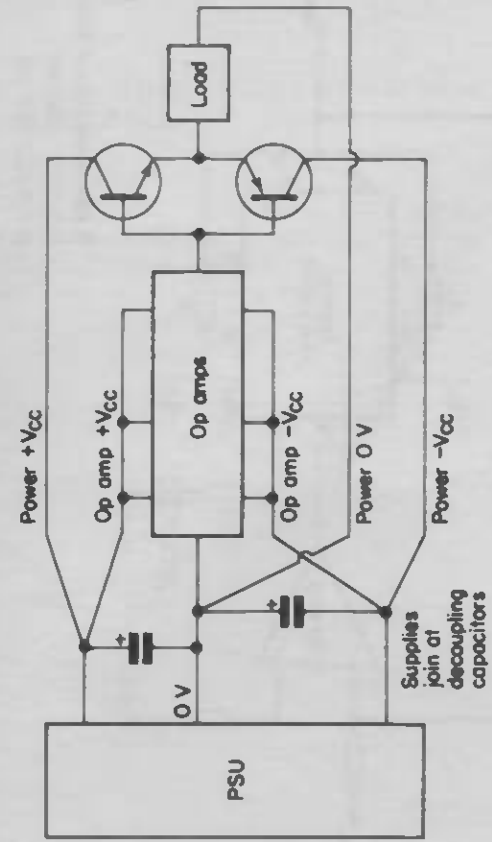


Fig. 2.25 Supply connections for heavy loads

giving a dissipation of nearly 2 watts. In most cases heat sinks of some sort will be necessary.

If high currents are used, care should be taken to use sensible return wires, as shown on Fig.2.25. Do not take several amps down Vero Board tracks! In extreme cases, separate power and Op Amp supplies can be used.

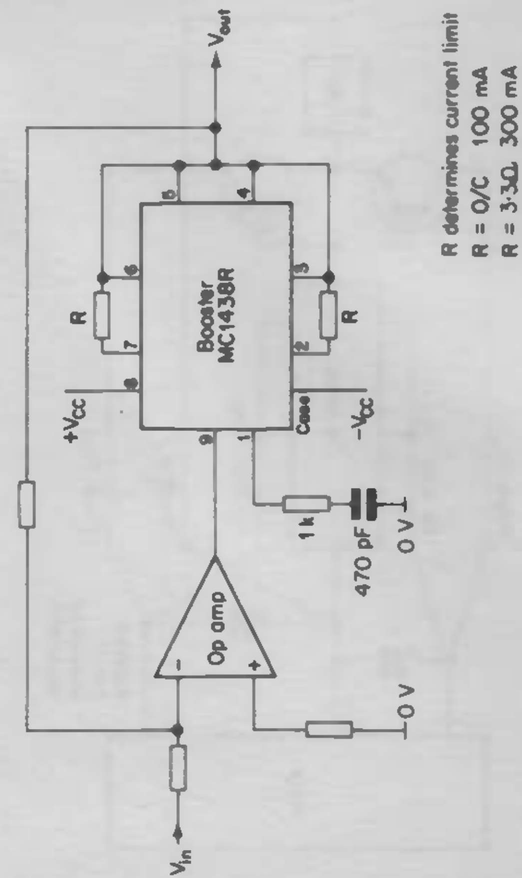


Fig. 2.26 Current booster I.C.P.

Where moderate currents are needed (up to 300mA) an elegant, although slightly expensive, solution is to use the current booster IC available from several manufacturers (MC1438R). This is connected as Fig.2.26 and gives increased current output for little trouble. The earlier comments regarding feedback still apply and feedback should still be taken from the load side of the booster IC.

2.10 POSITIVE/NEGATIVE AMPLIFIER

The circuit in Fig.2.27 is an ingenious combination of an inverting amplifier and a non-inverting amplifier. With the contact closed, the amplifier is identical to the inverting amplifier of section 2.2 with unity gain.

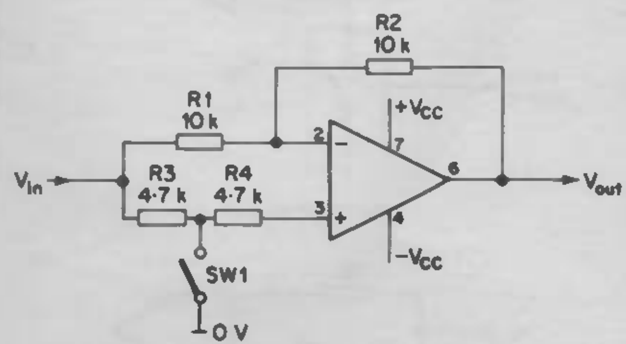


Fig. 2.27 Inverting/non inverting amplifier

With the contact open, the amplifier effectively becomes a voltage follower and is now a non-inverting amplifier with unity gain.

The contact in Fig.2.27 can be a simple relay contact or switch, but electronic contacts can also be used. In Fig.2.28a an FET is used. With the gate positive the FET exhibits a low

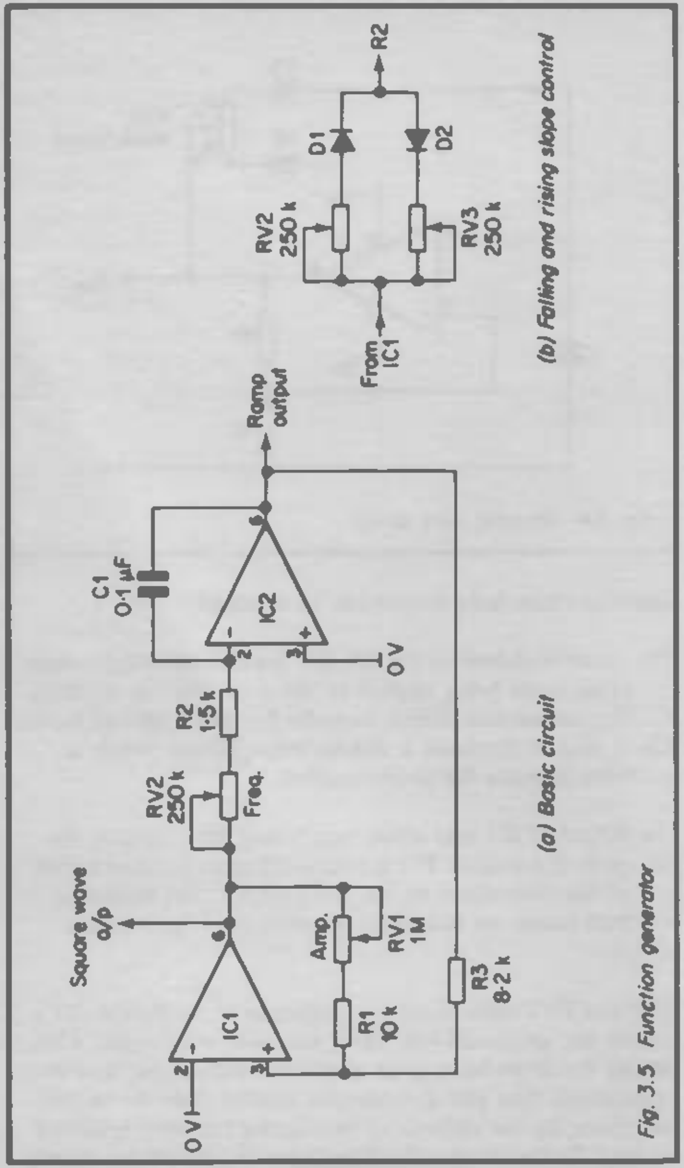


Fig. 3.5 Function generator

3.4 VOLTAGE CONTROLLED OSCILLATOR

An elegant voltage controlled oscillator can be made combining the oscillator from the previous section with the positive/negative amplifier of section 2.10. The circuit is shown on Fig.3.6. IC2 is an integrator and IC3 a Schmitt trigger as described in the previous section. The upper and lower trigger points of the Schmitt trigger are set by the zener diodes ZD1, ZD2. (see section 2.8).

IC1 is connected as a positive/negative amplifier with the control voltage applied as an input, and the sign determined by the field effect transistor FET1. The output of IC3 controls FET1, hence the output of IC1 will be either $+V_{in}$, or $-V_{in}$ as determined by the output signal.

The output of IC1 is applied to the integrator IC2. The circuit thus forms an oscillator similar to section 3.3. The larger the input voltage, the faster will the output of IC2 rise or fall, and hence the frequency of oscillation will increase. The relationship between frequency and input voltage is, in fact, very linear and the circuit will operate over a range from 50Hz to 10kHz with the values shown.

If the circuit is to be operated with values of V_{in} below 50mV, the input voltage offset should be balanced out by the addition of RV1. If the circuit is to be operated at high frequencies, amplifiers with high slew rates should be used for IC1 and IC3.

3.5 WIEN BRIDGE OSCILLATORS

3.5.1 Introduction

Sine Wave Oscillators with any reasonable degree of purity are rather difficult to make. For oscillation to occur, the total phase shift round the amplifier must be 360 degrees and the gain exactly unity. The first criteria is not difficult to achieve, but the second is usually achieved by allowing than the value of ZD1, C4 will be discharged by R5, and FET1 will present a low resistance between source and drain. If the output amplitude exceeds the value of ZD1, C4 will charge negatively, and FET1 will increase the resistance between source and drain. This in turn reduces the amplifier gain and the output amplitude. The circuit thus stabilises with the peak to peak output voltage twice the value of ZD1.

3.6 QUADRATURE OSCILLATOR

All integrators turn a sine wave into a cosine wave. This is equivalent to a phase advance of 90°. The low pass filter of section 5.2.2 introduces a phase delay of 90°. It follows that an integrator and low pass filter connected with feedback will oscillate at the cut off frequency of the filter when the closed loop gain is unity.

Two practical circuits are shown on Fig.3.10, one for low frequency and one for higher frequency applications. Somewhat interestingly the oscillation frequency is not affected by the values of R3 and C3, providing the loop gain is greater than unity. Zeners ZD1 and ZD2 act as limiters to reduce the loop gain by clipping the peaks. The low pass filter of IC1 and its associated components effectively remove the distortion giving very clean sine and cosine outputs.

3.7 CRYSTAL OSCILLATOR

A crystal controlled square wave oscillator can be made by using a crystal in the shunt resonant mode and an amplifier. The circuit is shown on Fig.3.11. The circuit oscillates at the frequency where the crystal exhibits minimum impedance. The crystal thus works at its fundamental frequency. It should be noted that most RF crystals nominally operate at some overtone frequency and will operate at some fraction of their marked frequency in this circuit.

The amplifier must have high frequency gain and high slew

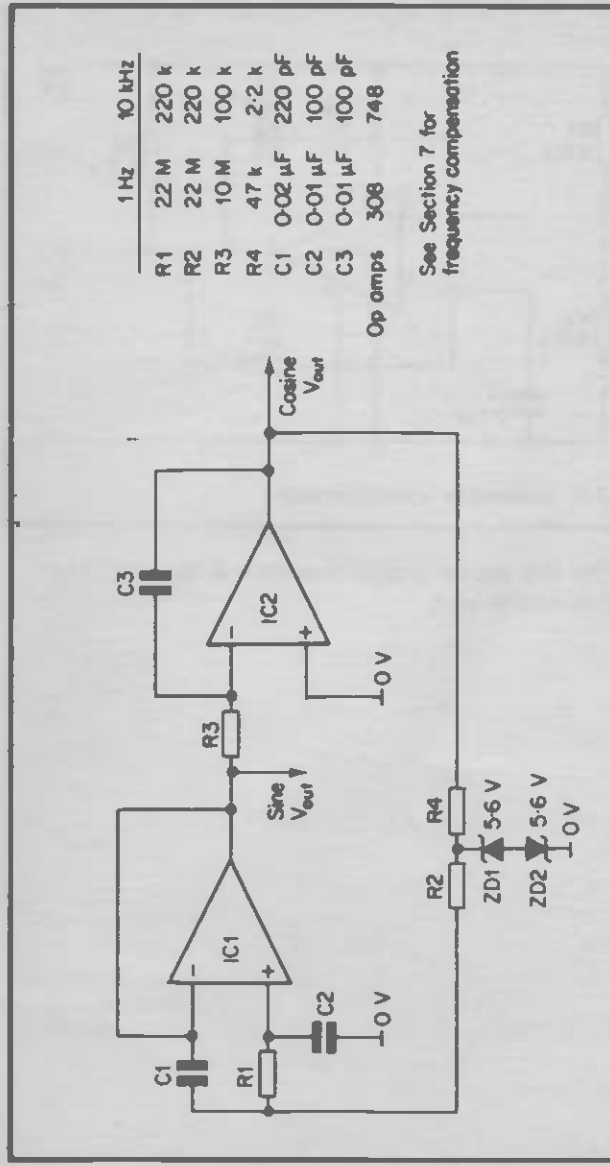


Fig. 3.10 Quadrature oscillator

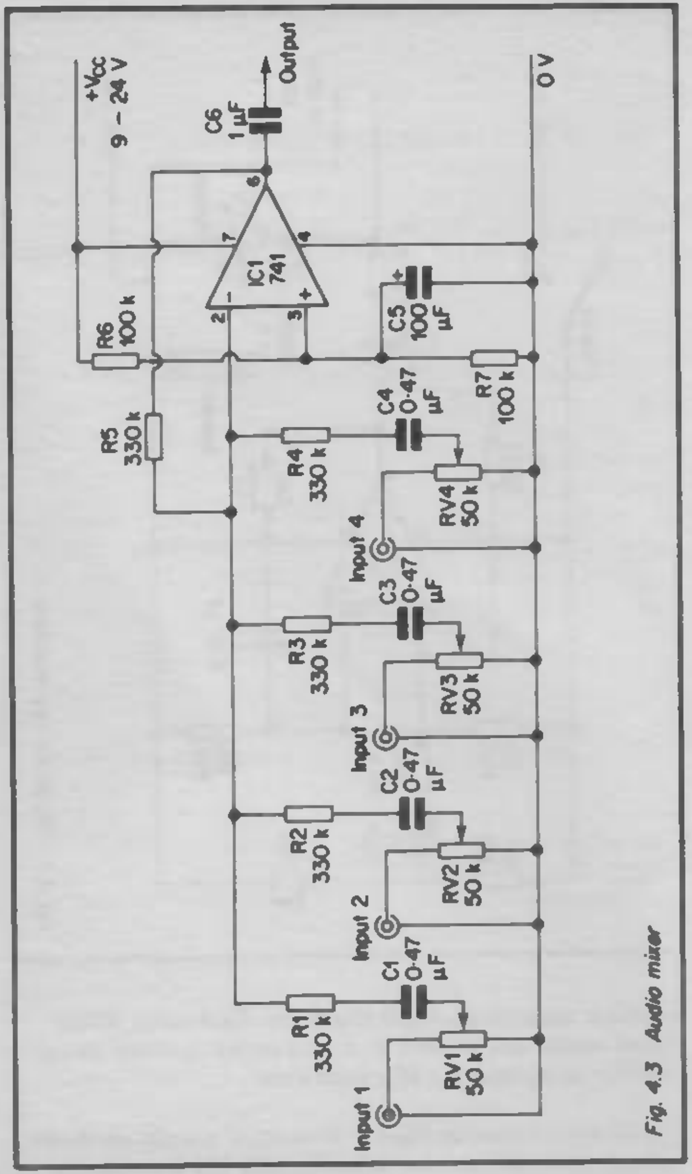


Fig. 4.3 Audio mixer

as an adder with the junction of R1 – R4 forming a virtual earth. The output voltage is thus the sum of the signals at the sliders of RV1 – RV4.

RV1 – RV4 set the individual levels. For a really professional looking unit slider controls should be used. R5 could be made a variable overall gain control if required. With the values shown the circuit has unity gain, but R1 – R4 could have different values if the three input signals have widely different levels.

R6 and R7 bias the circuit midway between the supply rails, and C5 decouples any noise on the supply. C1 to C4 AC couple the input signals to the input resistors, and prevent DC current flowing through the wipers on the input potentiometers.

The 741 and 709 are rather noisy devices, and should be avoided if the mixing is done with low level signals. Special audio operational amplifiers are available at slight expense which have very low noise levels, (e.g. 748).

4.4 POWER AMPLIFIER

The design of power amplifiers is a rather specialised art if Hi-Fi quality is required, but a simple power amplifier of reasonable quality can be made from an Op Amp and a few additional components.

The circuit is shown on Fig.4.4 and is basically the high current stage described in section 2.9, re-arranged slightly for AC operation. The gain of the stage is determined by R1 and R2 and should be chosen such that the maximum input signal gives an output signal just inside the supply rails. If, for example, the input signal is 0.5 volts peak to peak, and we have a 24 volt supply rail, the maximum gain will be about 20/0.5. i.e. 40 times and R1 and R2 should be chosen accordingly, 10k for R1 and 390k for R2 being suitable values. It should be remembered that the input signal peak to peak value priced record decks, and give acceptable, but not high fidelity signals. The output from these cartridges is a high impedance signal of around 250mV amplitude.

d. Radio Tuner. Almost all tuners give an output with a flat response and an amplitude of about 100mV.
e. Tape Decks. Although the output from a tape head requires considerable equalisation, this is normally done within the tape deck itself. The signal from a tape deck is therefore similar to that from a tuner with flat response and 100mV amplitude.

A pre-amplifier thus has to deal with a wide range of signal sources of greatly different characteristics. The use of negative feedback with an operational amplifier allows a simple solution to the problem.

A useful pre-amplifier circuit is shown on Fig.4.5. The heart of the circuit is 748 amplifier IC1. The 748 is used in preference to the 741 because of its better frequency response and lower noise. For experimental purposes a 741 would suffice, however.

The amplifier is connected in its non-inverting mode (which gives lower noise, see section 2.7) with R5, R6 and R7 biasing the amplifier for unity DC gain. Offset effects are therefore negligible. Capacitor C2, however, acts as a low impedance to AC signals so the amplifier gain without the feedback via SW1b would be very high.

The circuit as drawn has 5 input sources for demonstration purposes. In practice the user would build the circuit with fewer inputs relevant to his own equipment. The coupled switches SW1a and SW1b select the source and feedback necessary to give the correct gain and response.

The tuner and tape inputs are simplest, so these will be dealt with first. These are selected on switch positions 4 and 5. The response is flat, so SW1b selects R11 as feedback. The gain is determined by the ratio of R11 and R6, C2 appearing as a low impedance to audio frequencies. With the value shown the gain is about 2 times.

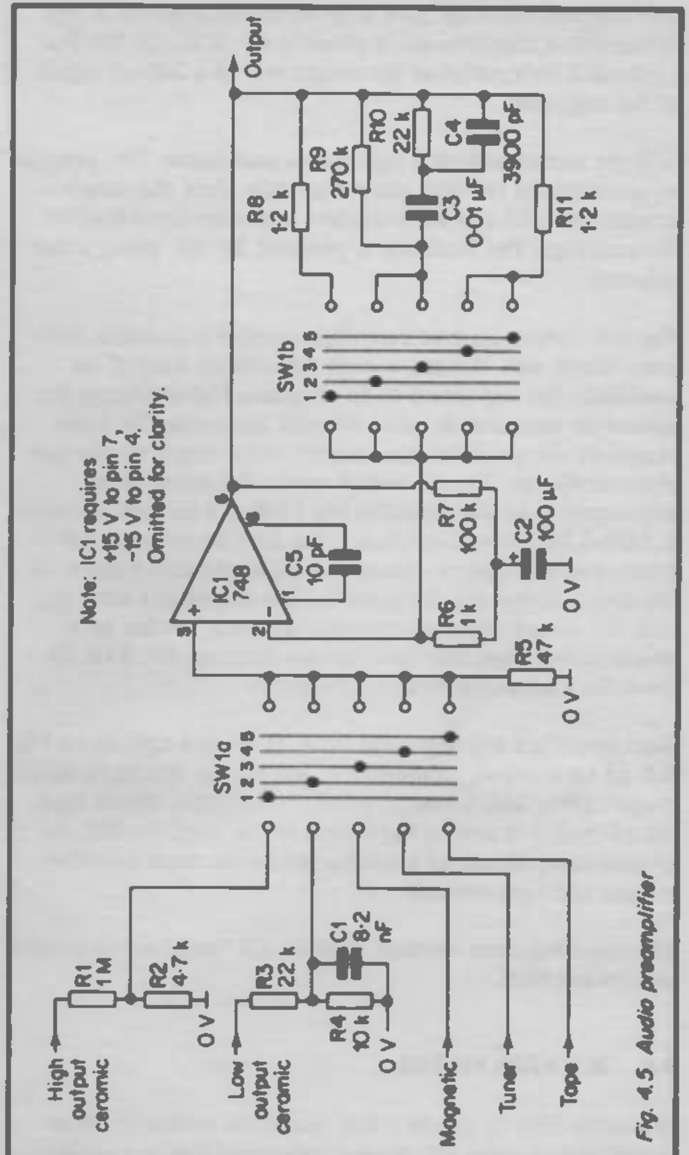


Fig. 4.5 Audio preamplifier

turntable. Compared with the belt drive used on Hi-Fi decks, direct drive can cause the motor rumble to be coupled into the arm. This rumble can be removed to a large extent by the use of a filter similar to Fig.4.7.

This is similar to the high pass filters described in section 5.3. With the values shown the circuit rejects all frequencies below 50Hz.

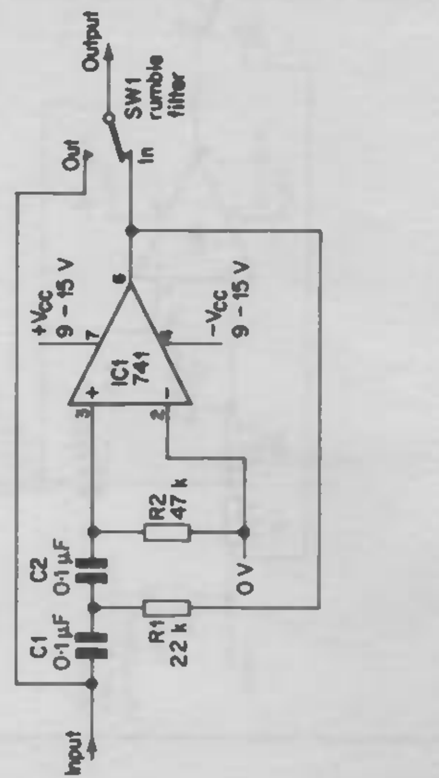


Fig. 4.7 Rumble filter

4.8 TONE CONTROL

Negative feedback allows the gain of an amplifier to be defined precisely. It follows that a very effective tone control, with separate treble and bass adjustment, can be made by using frequency dependant networks in the feedback of an amplifier.

There are many possible circuits (in fact it would be possible to fill a book with tone controls!) but one of the commonest

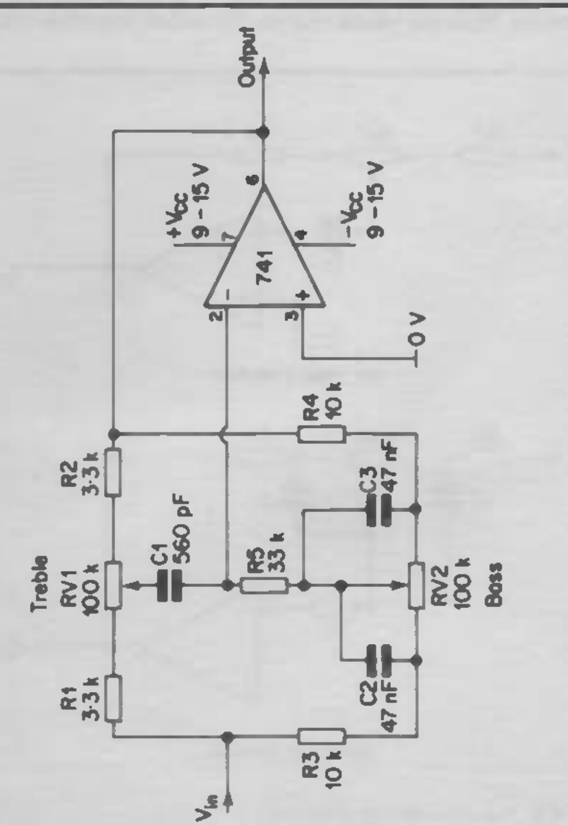


Fig. 4.8 Tone control

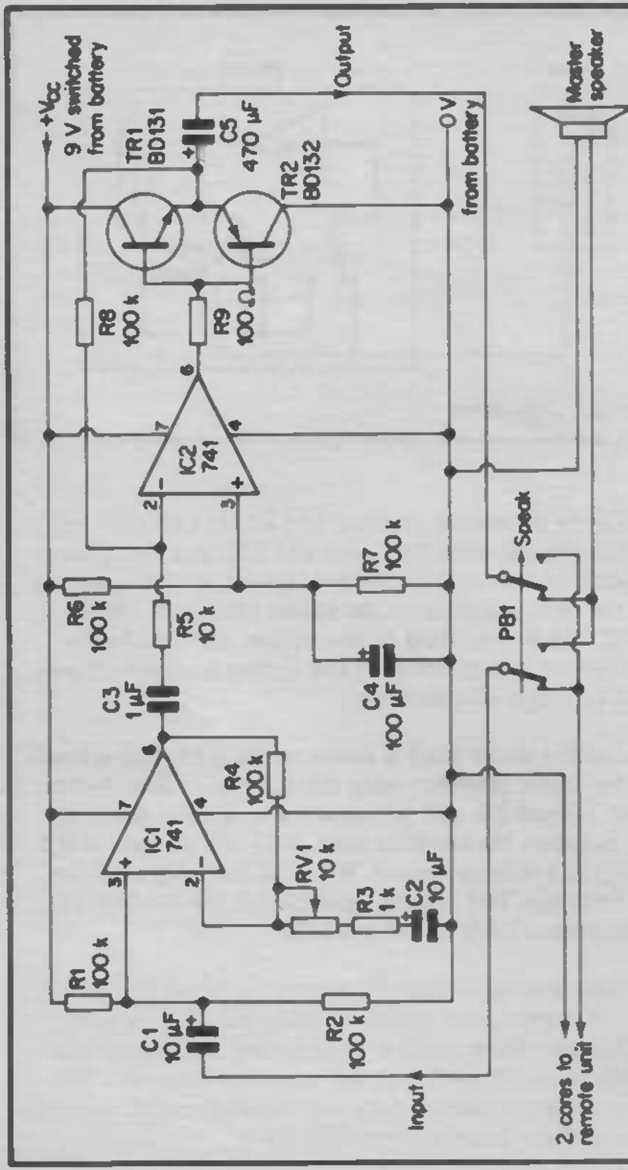


Fig. 4.14 Audio amplifier for baby alarm and telephone monitor

4.11 TELEPHONE MONITOR

A telephone monitor allows many people to listen to a telephone conversation. Post Office regulations forbid connection to telephone cables, but inductive pickup units are available from electronic hobbyshops. These consist of a multiturn coil attached to a rubber sucker. These can be stuck to the phone and pick up speech by electromagnetic induction. The best position for the sticker seems to vary from phone to phone, but the best place is usually adjacent to the handset speaker or on the side of the phone base.

The inductive sucker is connected to the input of the amplifier of Fig.4.14, from the previous section with a cheap loudspeaker connected to the amplifier output. In use, care must be taken in the setting of the volume control and the positioning of the unit to prevent the telephone microphone picking up the amplified speech. If this occurs a banshee wail comes from both the phone and monitor!

4.12 NOISE GENERATOR

Noise is something that designers of audio circuits normally try to avoid. A noise generator is, however, needed as the basis of many sound effects (steam engines, whistles, wind and rain etc.). The circuit shown on Fig.4.15 is a simple, but effective, noise generator.

The noise is generated by TR1. The base emitter junctions of a transistor acts as a zener diode when back biased above 5 volts. If the biasing current is very low, the resulting zener voltage is very noisy.

The noise from TR1 is AC coupled to IC1 connected as a high gain amplifier giving noise of about 2 volts amplitude at the amplifier output.

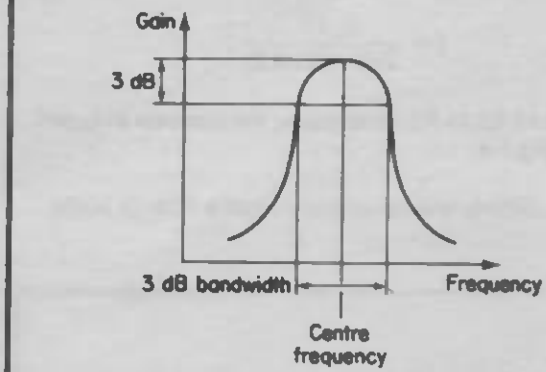


Fig. 5.7 The bandpass filter

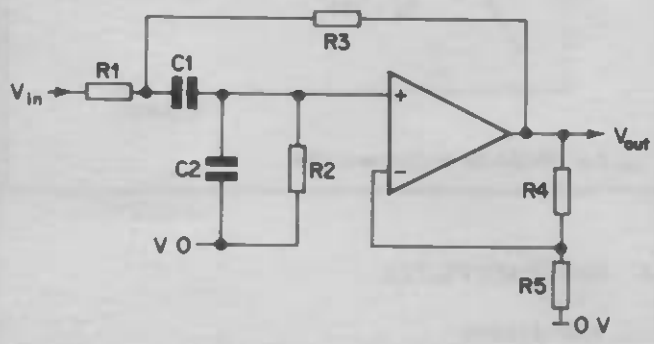


Fig. 5.8 Single amplifier bandpass filter; first circuit

simplify calculation R1, R2, R3 should be equal (denoted by R) and C1 and C2 should be equal (denoted by C). The centre frequency, fo, is then given by:–

$$fo = \frac{\sqrt{2}}{2\pi RC}$$

The Q of the circuit is determined by R4 and R5 with:–

$$Q = \frac{R5\sqrt{2}}{4R5 - R4}.$$

An alternative single amplifier circuit is shown on Fig.5.9. The components should be chosen such that R1 and R2 should be equal (denoted by R) and R3 should be chosen to be 2R. C1 and C2 should be made equal (denoted by C). The equations are somewhat simpler, with:–

$$fo = \frac{1}{2\pi RC}$$

and

$$Q = \frac{R5}{2R5 - R4}$$

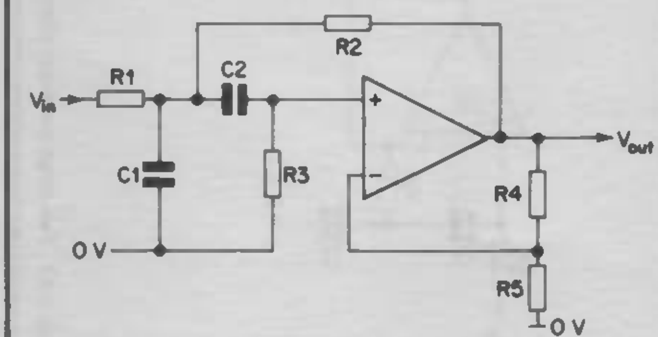


Fig. 5.9 Single amplifier bandpass filter; second circuit

5.4.3 Two Stage Bandpass Filter

The circuit of Fig.5.10 is easier to comprehend than those of the previous section, although it uses two amplifiers. IC1 is a low pass filter similar to section 5.2.1, and IC2 a high pass filter similar in principle to section 5.3.1. The centre frequency is determined by

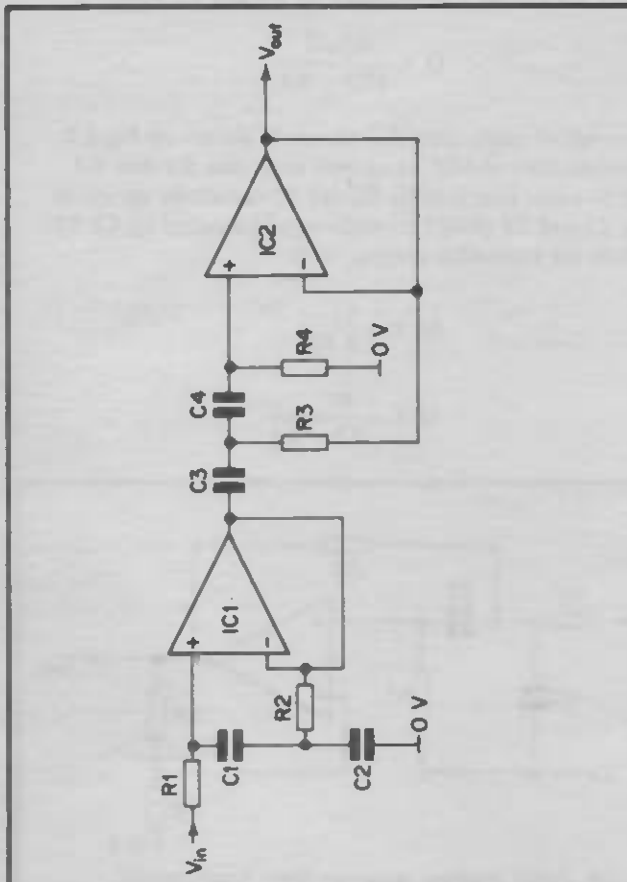


Fig. 5.10 Two stage bandpass filter

$$fo = \frac{1}{2 \pi RC}$$

if R1 = R2 = R3 = R4 = R and C1 = C2 = C3 = C4 = C.

The bandwidth is determined by the cutoff frequencies as described in sections 5.2.1 and 5.3.1. If the above equalities do not hold, the upper and lower frequency cutoff can be determined separately.

5.5 NOTCH FILTERS

5.5.1 Introduction

A notch filter is the opposite of a bandpass filter in that it rejects a band of frequencies. Commonly the notch filter is used to reject 50Hz mains hum (60Hz outside Great Britain) in sensitive audio circuits and measuring instruments. The centre frequency and Q of a notch filter are defined in a similar manner to those of a bandpass filter.

5.5.2 Single Amplifier Notch Filter

The circuit of Fig.5.11 gives a notch filter of very high Q. To simplify the design, the values should be chosen such that:–

$$\begin{aligned} R1 &= R2 = R \\ R3 &= R/2 \\ C1 &= C2 = C \\ C3 &= 2C \end{aligned}$$

If the above conditions are met, the centre frequency is given by:–

$$fo = \frac{1}{2 \pi RC}$$

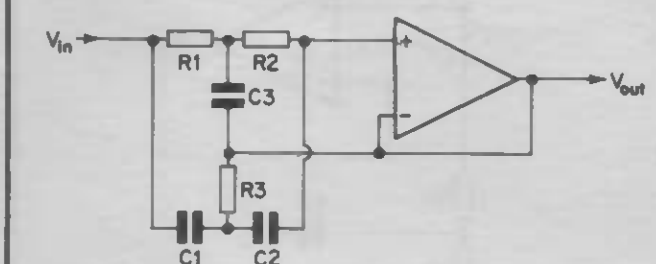


Fig. 5.11 High Q notch filter

5.5.3 Two Amplifier Tunable Notch Filter

The circuit of section 5.5.2 gives a short notch, but the equalities necessary make it difficult to tune. The circuit of Fig.5.12 can be tuned by a single variable capacitor, C1. Normally C1 will be a few hundred picafarads and C2 several microfarads. As usual, the design is simplified by the resistor equalities below:–

$$R1 = R2 = R3 = R$$
$$R4 = R5 = R/2$$

The centre frequency is then given by:–

$$fo = \frac{1}{\pi R \sqrt{C1C2}}$$

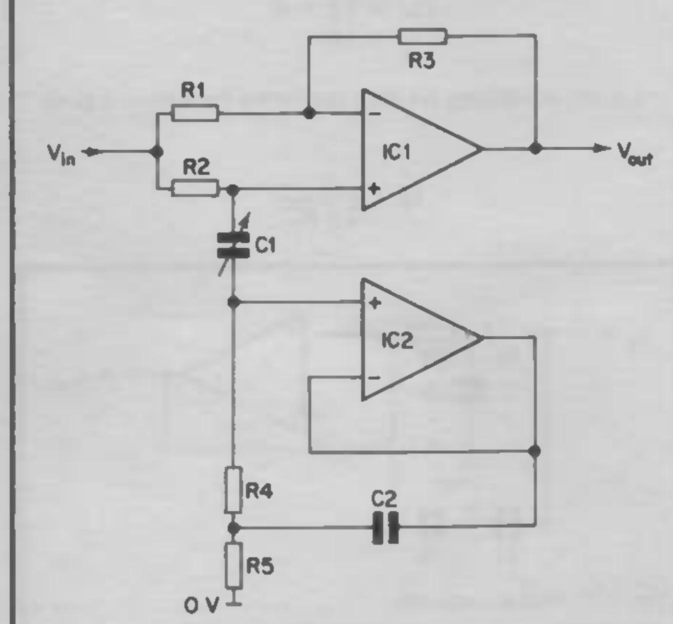


Fig. 5.12 Tunable notch filter

5.5.4 Adjustable Q Notch Filter

The circuit of Fig. 5.13 allows the Q of a notch filter to be varied by a single potentiometer without varying the centre frequency. The potentiometer can be any reasonable value, the Q of the circuit being determined by the ratio R_a/R_b. As usual, some equalities must be observed:–

$$R1 = R2 = R$$
$$R3 = R/2$$
$$C1 = C2 = C$$
$$C3 = 2C$$

The centre frequency is given by:–

$$fo = \frac{1}{2\pi RC}$$

The adjustable Q notch filter is very useful in low level measuring instruments.

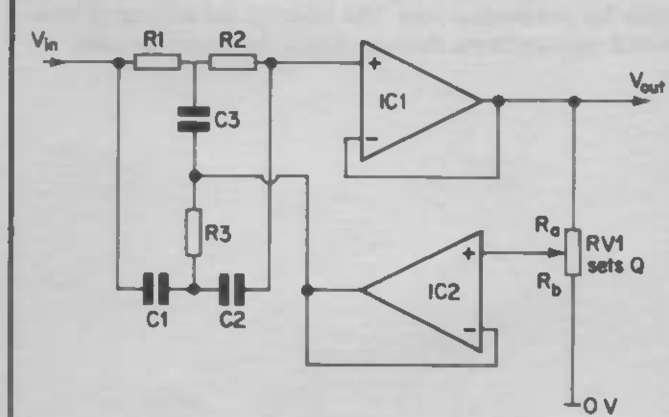


Fig. 5.13 Adjustable Q notch filter

5.6 PRACTICAL OBSERVATIONS

The obvious is sometimes overlooked, but it should be noted that in all the above equations resistors must be in OHMs and capacitors in FARADS giving results in Hz (cycles per Second for older readers!)

Where equalities are given, precision resistors (at worst 1% tolerance) should be used and close tolerance capacitors. Multiples and division by two is often needed in the equations, and this is best achieved by one value of resistor and capacitor throughout, and using parallel or series combinations to produce the multiples required.

It will often be found that none of the preferred value resistors or capacitors give the required frequency. Series combinations of resistors should be used in preference to variable resistors. This is inelegant, but one of the sad facts of life in filter design. Alternatively precision wire wound or thin film resistors can be ordered to specific values. Although prohibitively expensive for home "one-offs' this approach is economically viable for production runs. The inherent inductance of wire wound resistors limits their use to low frequency circuits.

CHAPTER 6
MISCELLANEOUS CIRCUITS

6.1 INTRODUCTION

The circuits in the previous sections fall into neat labels like, "Audio Circuits" or "Filters". Those described in this section do not give themselves to easy categorisation, but are none-theless very useful circuits demonstrating the versatility of the Operational Amplifier.

6.2 WINDOW DETECTOR

The window detector is a circuit that indicates if an input voltage is within specified limits. It can thus be used as part of an industrial control scheme, giving an alarm if the measured variable (temperature, liquid level, etc.) goes outside the correct levels. Alternatively it can be used as part of an automatic test box for anything from resistors to power supplies.

The circuit shown on Fig.6.1a has the response of Fig.6.1b. With the input voltage between V_1 and V_2 the output relay is energised. If the input voltage goes above V_1 or below V_2 the relay de-energises. V_1 and V_2 thus determine the "correct" range and can be any value or polarity (except, of course, the obvious one that V_1 must be more positive than V_2!)

Amplifier IC1 is a simple inverting amplifier, to allow small differences to be detected with reasonable values of V_1 and V_2. The gain is adjusted by RV1. If required an offset can be added by RV2, if it is required to have a small band on top of a large voltage (e.g. with a correct range 5V to 5.5V, RV2 could be used to null out 5V, allowing a gain of 20 to be used on IC1 to amplify the 0.5V band to 10 volts at IC1 output).

The output of IC1 is applied to the non-inverting input of IC2 and the inverting input of IC3. These amplifiers have no feed-

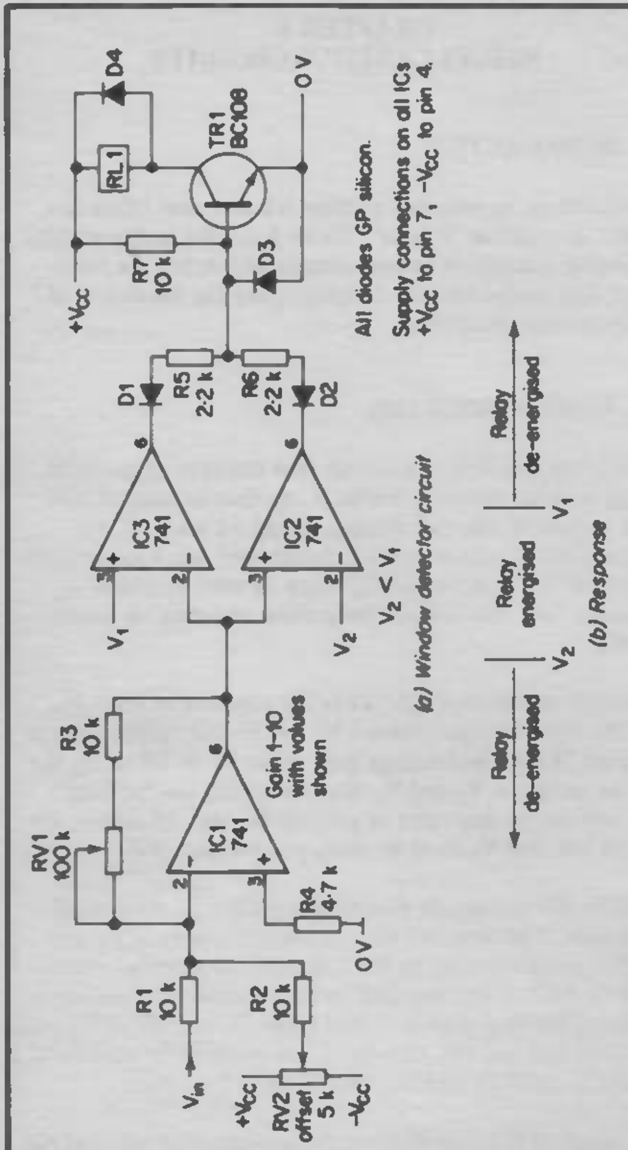


Fig. 6.1 Window detector

back, and hence operate with full gain. For all practical considerations, therefore, the outputs will be at either the positive or negative supplies depending on the comparison of IC1 output with V_1 or V_2.

The outputs of IC2 and IC3 are connected to the base of TR1. With the output of IC1 between V_1 and V_2, both outputs will be positive and TR1 will be turned on with base current supplied by R7. If the output of IC1 goes more positive than V_1 or more negative than V_2 the corresponding output will go to the negative supply rail, and turning TR1 off via R5 or R6. Relay RL1 will thus de-energise if the input goes outside the predetermined range.

Possible sources for V_1 and V_2 are shown on Fig.6.2. Note that because IC1 is connected as an inverting amplifier (to allow an offset to be added), voltage V_2 corresponds to the maximum trigger voltage at R1, and V_1 to the minimum trigger voltage. Where the supply rails are of dubious stability (e.g. a battery supply) the zener diode sources are preferable.

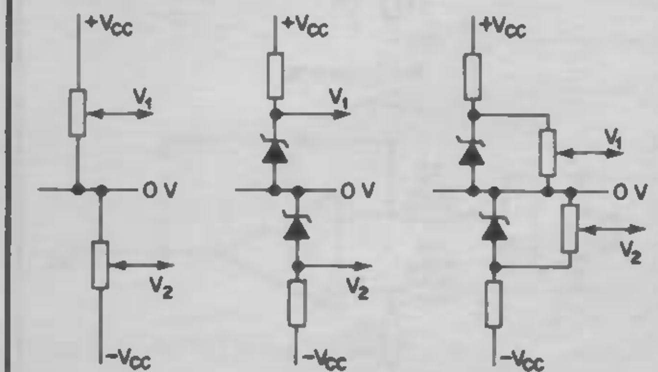


Fig. 6.2 Voltage references for window detector

6.3 VOLTAGE TO CURRENT AND CURRENT TO VOLTAGE CONVERSION

If a voltage is to be conveyed down a cable for any distance it is desirable to convert it to a current and reconvert it a voltage at the receiving end because of the superior noise rejection given by a current loop. In addition many industrial actuators and instruments require a current input.

The circuit in Fig.6.3a has a voltage input on the non-inverting input of IC1. The load is connected between the amplifier output and the inverting input, and thence to ground via R1.

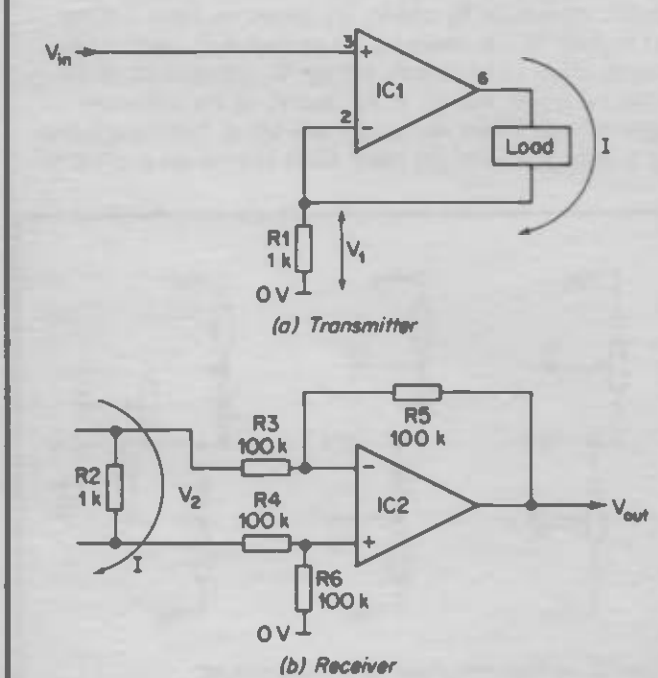


Fig. 6.3 Voltage/current/voltage conversion

To maintain balance the amplifier output will drive a current, I, through the load and R1 producing a voltage at the inverting input of I.R1.

Since $V_1 = V_{in}$

$$I = \frac{V_{in}}{R1} \quad \text{which is totally independent of the load.}$$

The circuit outputs a current into the line which is solely dependent on the input signal, and independent of the load.

At the receiving end, the current can be converted to a voltage, if required, by the circuit of Fig.6.3b. The current passes through resistor R2 and produces a voltage V_2. This voltage is connected to a differential amplifier (see section 2.5) to remove any common mode noise, giving the output voltage V_{out}.

If R1 = R2 and R3 – R6 are equal, the whole circuit has unity gain and $V_{out} = V_{in}$.

6.4 RAMP CIRCUIT

In motor drive circuits the acceleration is often required to be limited to prevent excessive currents or mechanical strain. The low pass filter of section 5.2.1 will give an exponential response, but a more elegant solution is shown on Fig.6.4. This has a constant rate of change of output until $V_{out} = -V_{in}$. The response can thus be summarised as Fig.6.5.

On Fig.6.4, IC1 is used as a comparator, comparing V_{out} with V_{in}. If the non-inverting input of IC1 is positive, then the output of IC1 will be at the positive supply rail. Similarly if the non-inverting input is negative the output of IC1 will be at the negative supply rail.

R1 and R2 determine the gain of the circuit, and for unity gain are equal. IC1 will thus balance when $V_{out} = -V_{in}$.

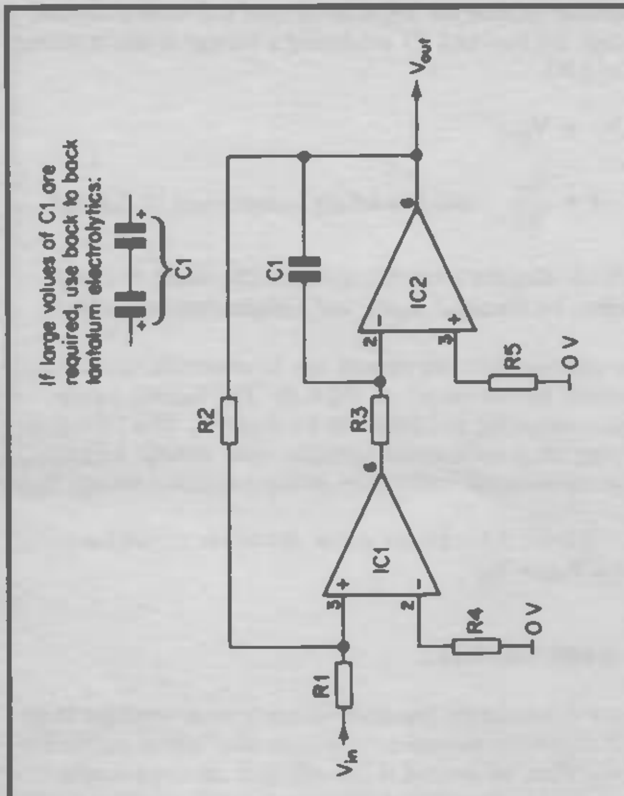


Fig. 6.4 Ramp circuit

IC2 acts as an integrator, except the input to R3 can only be the positive supply or the negative supply. Suppose V_{in} is greater than $-V_{out}$. The input to R3 will be the positive supply, and V_{out} will change in a linear manner until V_{out} again equals $-V_{in}$. Similarly, if V_{in} is less than $-V_{out}$, we will get a linear change of V_{out} until balance is attained again. When V_{out} equals $-V_{in}$ the output of IC1 is nominally 0V, but in practice will dither randomly compensating for offset currents in the integrator.

The gain of the circuit is simply $-R2/R1$.

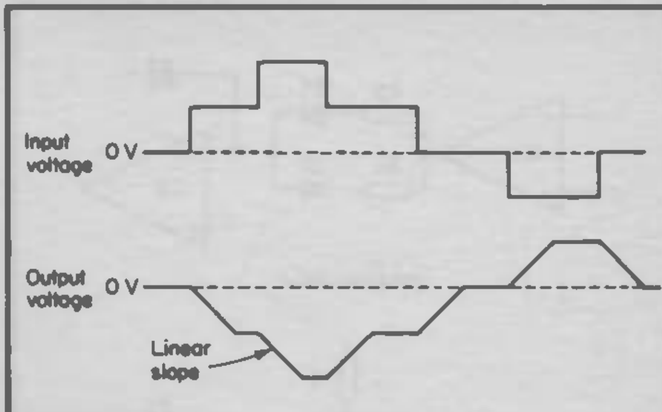


Fig. 6.5 Response of ramp circuit

The rate of change of the output is given by

$$\frac{V}{R3.C1} \quad \text{volts/second,}$$

where V is the supply rail voltages (assumed equal ±V) and R is expressed in ohms and C in farads. Alternatively R3 can be in megohms and C1 in microfarads.

If different rise and fall rates are required, either of the circuits of Fig.6.6a and b can be used. In each case R3 determines the negative output ramp rate, and R4 the positive rate, with the above equation still applicable. Fig.6.6b has the advantage that V_1 and V_2 need not be the supply rails, and could be derived from elsewhere.

6.5 PHASE ADVANCE CIRCUIT

The phase advance circuit is almost the opposite of the ramp circuit. It is used to "kick" an output before settling down to the final value as shown on Fig.6.7b. It is very useful for starting motors connected to loads with high inertia.

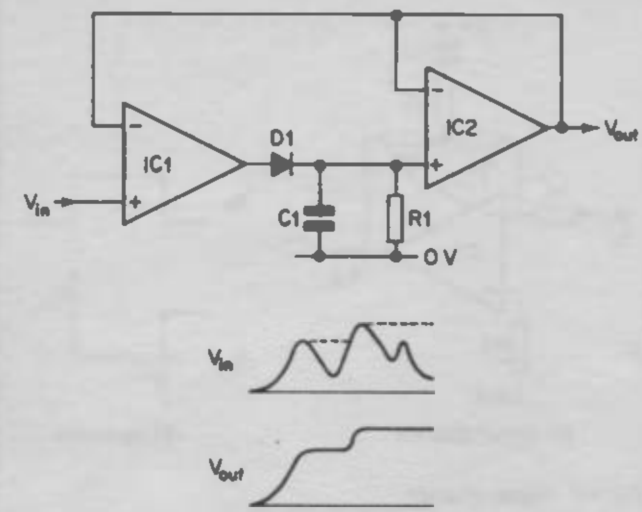


Fig. 6.8 Peak picker

mined by the time constant R1C1).

If D1 is reversed, the circuit stores the minimum (or most negative) input voltage.

6.7 SAMPLE AND HOLD

The sample and hold circuit is a variation on the peak picker of section 6.6. The sample and hold is used to "freeze" a varying signal by taking a snapshot picture. The resulting stable voltage can then be measured without ambiguity.

The basis is shown on Fig.6.9. With SW1 closed, the voltage on C1 and the output voltage will equal the input voltage as described in the previous section. If SW1 is open, C1 will hold the voltage at the instant the switch opened. In a sample and hold

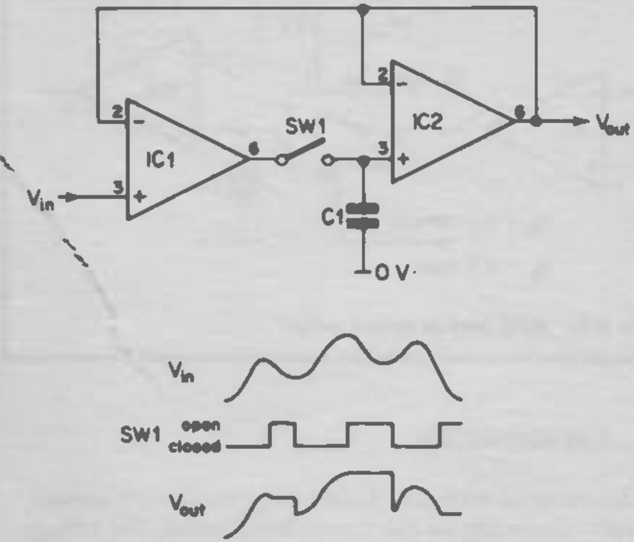


Fig. 6.9 Sample and hold

circuit, SW1 is normally open and is closed briefly to obtain a sample.

SW1 can be any form of switch. Reed relays were often used (and still are in some applications) but the solid state switches provided by FETs are ideally suited. Fig.6.10 shows a circuit using the popular 4016 CMOS switch. The circuit shown allows any voltage between 1 volt and 14 volts to be stored. More expensive switches are available (e.g. the 7502) which can be used with two supply rails.

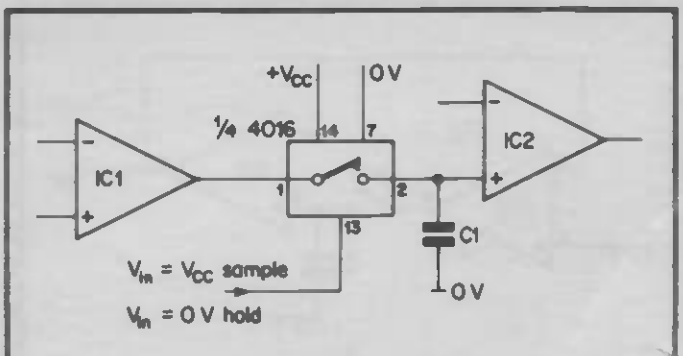


Fig. 6.10 4016 used as sample switch

6.8 THERMOMETER

The base-emitter voltage of a conducting transistor is around 0.5 volts, depending on the current being passed. The voltage, however, is also temperature dependent and changes by 2mV/°C for every transistor. This change can be made the basis of a very useful thermometer.

The circuit is shown on Fig.6.11. The temperature probe is TR1, biased into conduction by R1, R2 and RV1. The resulting base-emitter voltage is connected to the non-inverting input of IC1, which is connected as a non-inverting amplifier with gain adjustable around 50. The inverting input is biased at about 0.5 volts negative by R4, R3 and RV2.

RV2 acts, therefore, as a coarse zero control and RV1 as a fine zero control by varying the current through TR1. To set the zero, the probe should be immersed in melting ice and RV2 and RV1 adjusted for zero reading on the meter.

The probe should then be immersed in boiling water and RV3 set for 10 volts at the meter. The meter now reads 0 – 10 volts for 0 – 100°C.

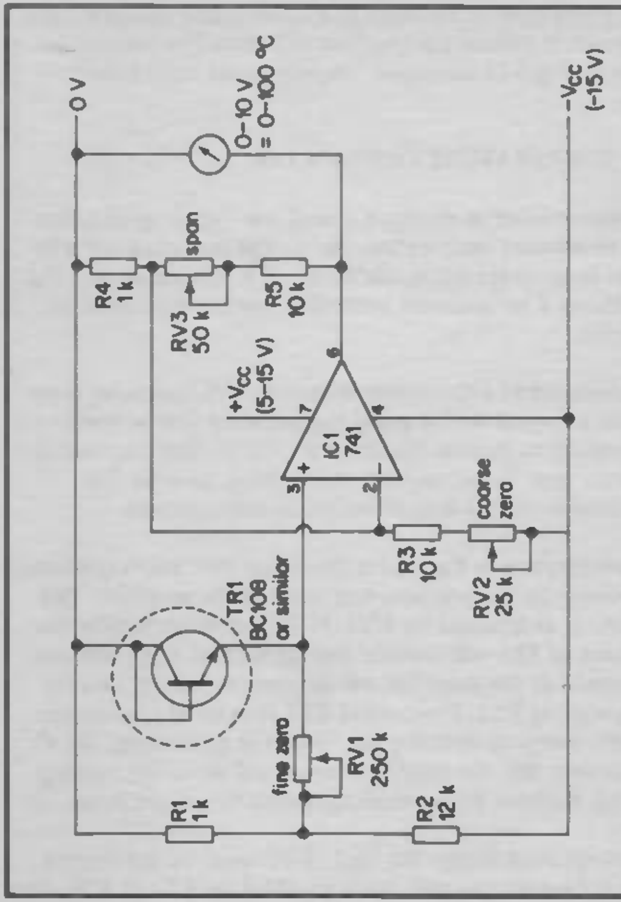


Fig. 6.11 Thermometer

The thermometer can be calibrated for any temperature range, but remember that transistors will be destroyed at temperatures in excess of 150°C.

The –15 volt rail must be well stabilised since the biasing for the non-inverting input is derived from this rail. The positive rail can be any value in the range 6 volts to 15 volts and need not be particularly well stabilised.

If the probe is to be mounted at a considerable distance from the circuit, a differential amplifier IC2 should be included as shown on Fig.6.12 to reduce common mode noise effects.

6.9 TEMPERATURE CONTROLLER

The thermometer in section 6.8 used the V_{BE} drop of a transistor to measure temperature. An alternative technique is to use the large temperature coefficient of a thermistor, and Fig. 6.13 shows a temperature controller using a thermistor as the sensor.

The resistance of a thermistor decreases with increasing temperature, a typical device going from around 10k at room temperature to around 100ohms at 100°C. The response is, however, non linear making them more suitable for temperature control than temperature measurement.

The temperature in Fig.6.13 is sensed by Th1, and is converted to a voltage on the non-inverting input of the amplifier. The set point is determined by RV1. If the temperature falls, the resistance of Th1 will increase and the voltage at R1 will rise. The output of the amplifier will go positive turning TR1 on and energising RL1. Contacts of RL1 turn on the heater, circulation pump or whatever the circuit is controlling. On a temperature fall, the amplifier output will go to 0V, turning TR1 off. Resistor R6 provides hysteresis to prevent jitter.

The comparison bridge R1, Th1, RV1 must be fed from a stable voltage source, and this is provided by R7 and ZD1. The Zener is deliberately chosen at 5.6 volts at this value has practically zero temperature coefficient itself.

Thermistors are available for many temperature ranges, and with care control to better than 2°C can be obtained.

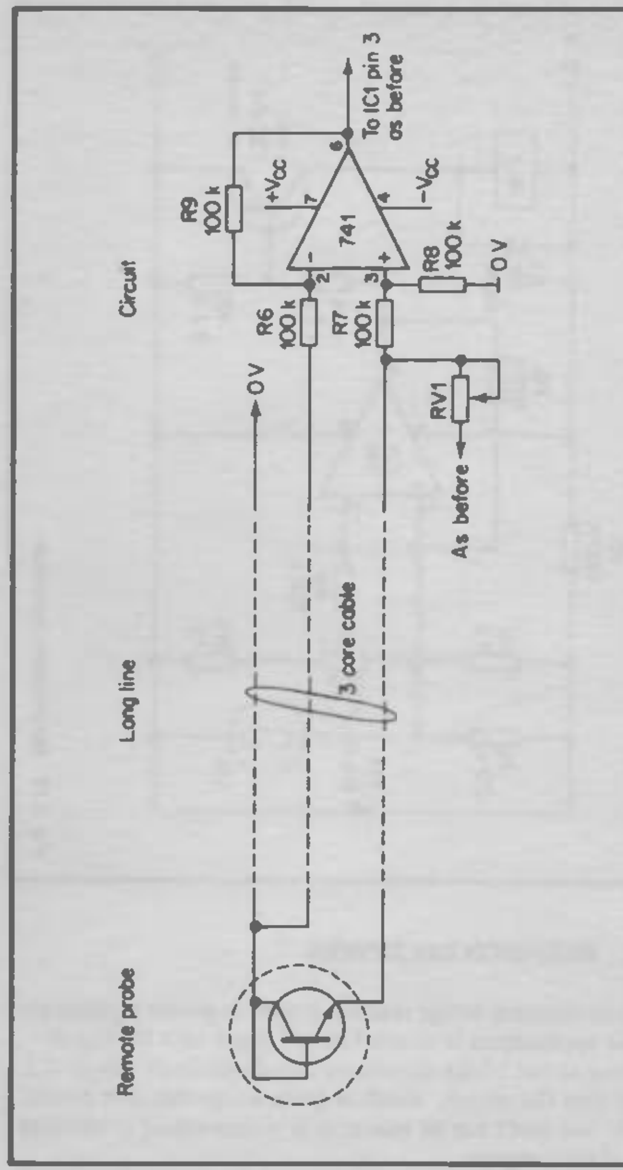


Fig. 6.12 Thermometer with remote sensor

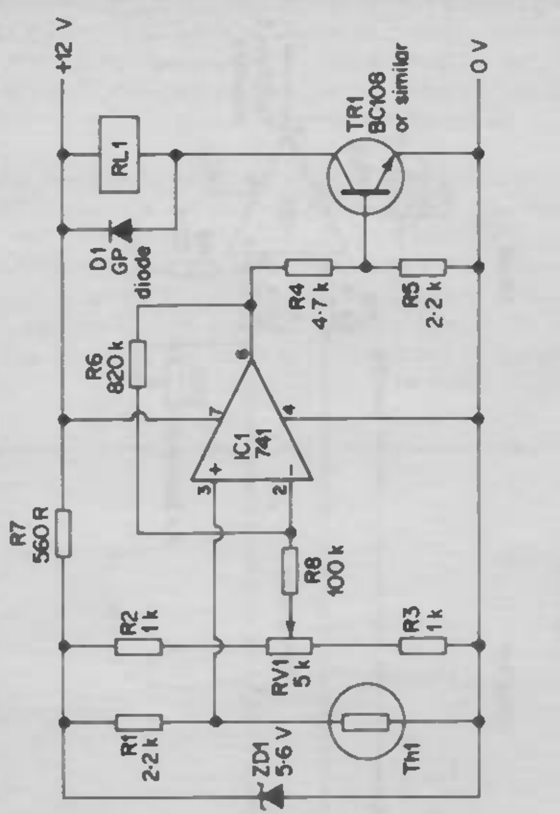


Fig. 6.13 Temperature controller

6.10 PRECISION RECTIFIERS

The conventional bridge rectifier is used in power supplies and similar applications to convert an AC signal to a DC signal. In doing so the bridge introduces two diode drops (about 1.5 volts) into the circuit, which is quite acceptable in a power supply but could not be tolerated in measurement or rectification of small signals.

The voltage drop across the diodes can be reduced by the open loop gain of an Operational Amplifier if the diodes are used in the amplifier feedback. There are many possible ways of obtaining perfect rectification, but the circuit described below is probably the easiest to understand.

First we must consider the halfwave rectifier of Fig.6.14. This is a conventional inverting amplifier with the inclusion of D1 and D2, and R1 and R2 equal in value. When V_{in} is on the negative half cycle, $V_{out} = -V_{in}$ by the arguments outlined in section 2.2. Note that the actual output of voltage of the amplifier, V_o, will be offset by the diode drop. During the negative cycle of the input we therefore get a positive half cycle output voltage with no voltage drop.

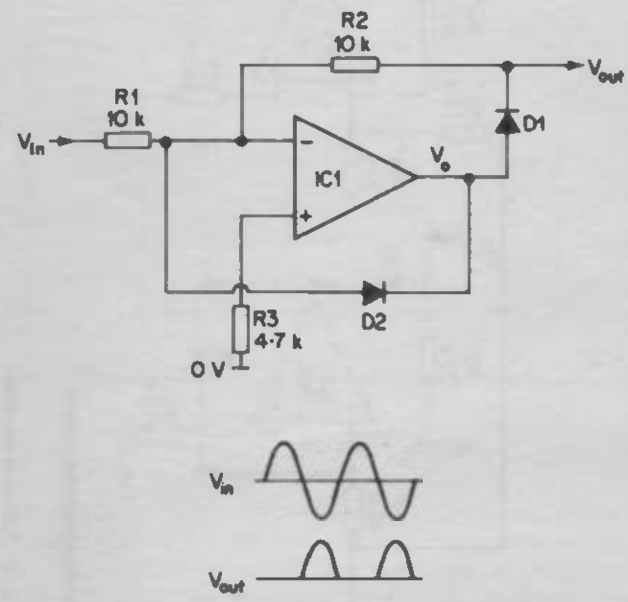


Fig. 6.14 Halfwave rectifier

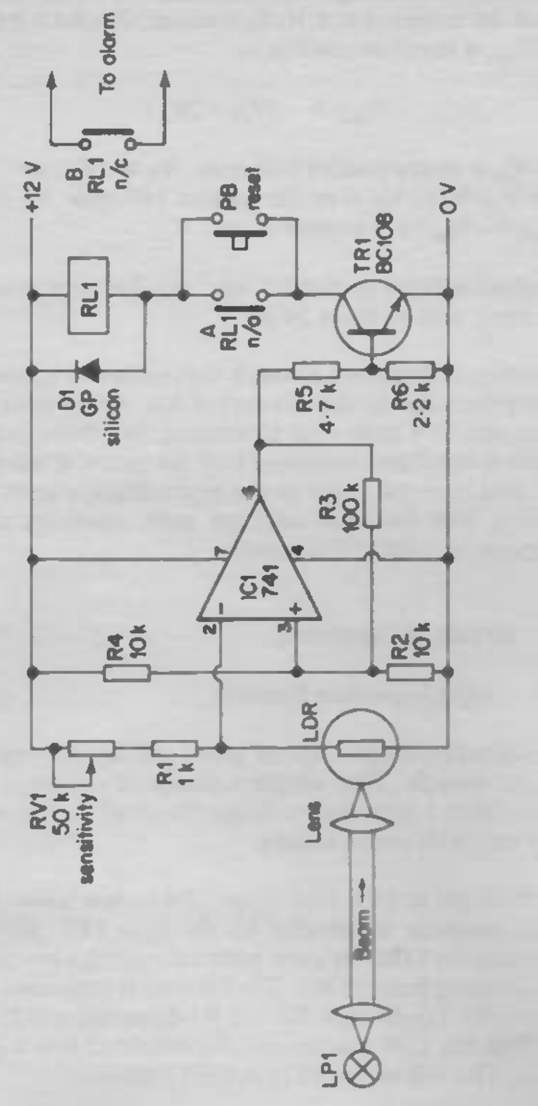


Fig. 6.16 LDR burglar alarm

the voltage at the inverting input rises above UTP, and IC1 output switches to 0V, turning TR1 off and de-energising RL1. Once dropped out, RL1 stays out because of contact A. Contact B sounds the alarm bell. Initially RL1 is energised by the arm push button PB1.

The variable resistor RV1 sets the circuit sensitivity. For maximum range, LDR and LP1 should be placed at the focus of a converging lens or a parabolic mirror (from a torch). If LDR is interchanged with R1 and RV1, the relay will energise when the beam is broken.

6.11.2 Photodiodes

Photodiodes are normal diodes operated with reverse bias. As usual a small leakage current flows, but in a photodiode this leakage current varies linearly with light. Fig.6.17 is a circuit to convert this leakage current change to a useful voltage which can be read on a meter or fed to a Schmitt Trigger similar to the preceeding section.

The non-inverting input of IC1 is a virtual earth, but the current input is derived from D1, not from a voltage and resistor as usual. To maintain the virtual earth, current flows through R1, and. –

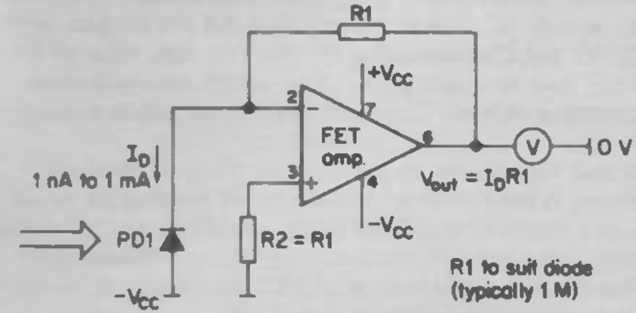


Fig. 6.17 Photometer

$$V_{out} = I_D R1.$$

The resistor R1 thus determines the gain of the circuit.

Because the currents involved are small (typically 1nA to 1mA) the amplifier needs to have low input bias current. The circuit works particularly well with FET input amplifiers.

6.11.3 Optical Link

The final optical sensor is the phototransistor. This is similar to the photodiode in that the leakage current of the transistor varies with incident light. The change in leakage current in the phototransistor is, however, far greater than that in the photodiode.

As a demonstration of the use of the phototransistor, Fig. 6.18a and b shows an experimental optical speech link. No one would seriously suggest this is any better than the field phones of section 4.2, but it is an interesting "fun" circuit.

Fig.6.18a is the transmitter, and is essentially a V to I converter, changing the voltage changes at the microphone to current changes through the Infra Red LED. For a 20mV signal, the current through the diode will change by 20mA.

Fig.6.18b is the receiver. This operates as a virtual earth amplifier similar to that in the preceeding section. C1 blocks DC, so only AC changes are amplified. R4 sets the gain, with R2, R3 and C3 maintaining DC bias (the high value of R4 would lead to unacceptably large offsets due to leakage currents on its own). C2 ensures that the AC gain is unaffected.

D1 and TR1 should be placed at the focus of parabolic mirrors. A hand torch is a useful case for building the circuit in, as a matter of fact. Glass lenses should be discarded as glass blocks infra red. For maximum sensitivity, a Wrattan 88A filter should be used in front of TR1. This blocks visible light, but passes infra red without attenuation.

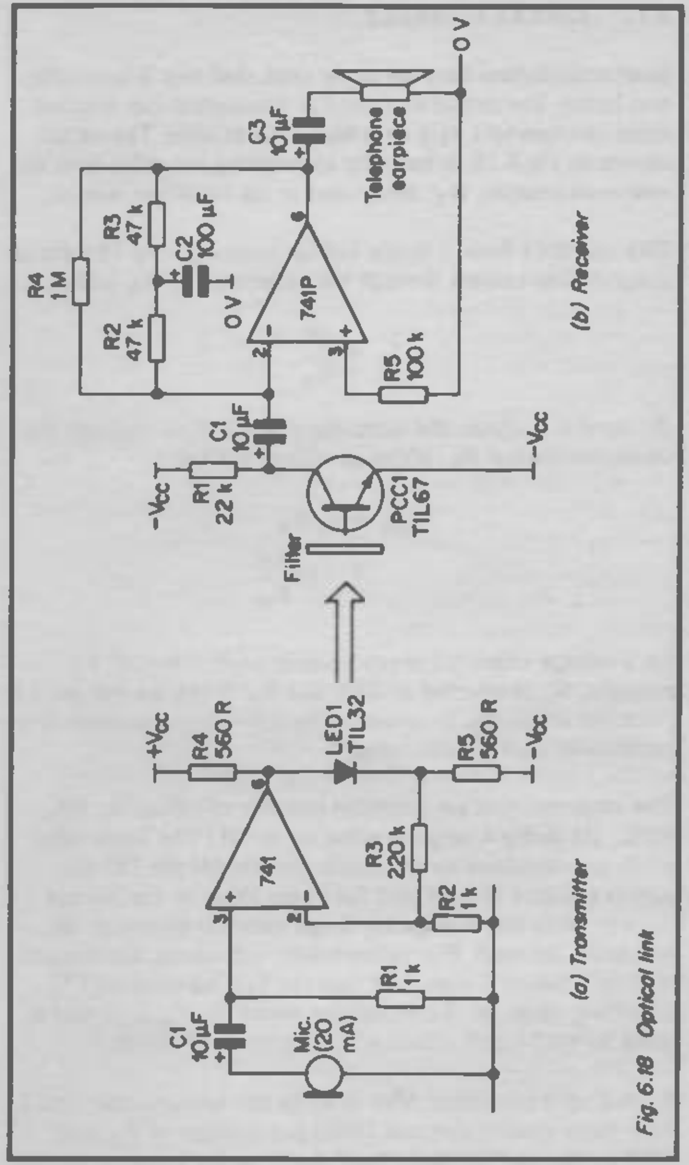


Fig. 6.18 Optical link

for zero on the meter.

RV1 sets full scale, and can be set with a multimeter or by using a precision resistor. With a multimeter RV1 is simply set to give 10 volts at point A. To set up with a precision resistor, the resistor is put in R_X and the correct range selected on SW1. RV1 is then adjusted to give the correct value on the meter. Once RV1 is set, it should be correct for all ranges.

6.13 LOW CURRENT METER

Most cheap multimeters have current scales that can only measure from a few milliamps upwards. The circuit described in this section extends the range of a multimeter down to below 10μA. The circuit is shown on Fig.6.20.

IC1 is an FET Op Amp (for low bias current) with a gain of 50 (set by R1 and R2) from V_{in} to V_{out}. The current through the meter is thus:–

$$I_{out} = 50.\frac{V_{in}}{R3}$$

The unknown current flows through the selected range resistor, chosen to give 5mV for full scale. For a selected range at full scale, therefore:–

$$I_{out} = \frac{50 \times 5 \times 10^{-3}}{250} \text{ A}$$

$$= 1\text{mA}$$

A 1mA full scale meter will therefore read correctly on each range.

RV1 sets the meter to zero, and D1, D2 are used to protect against high currents.

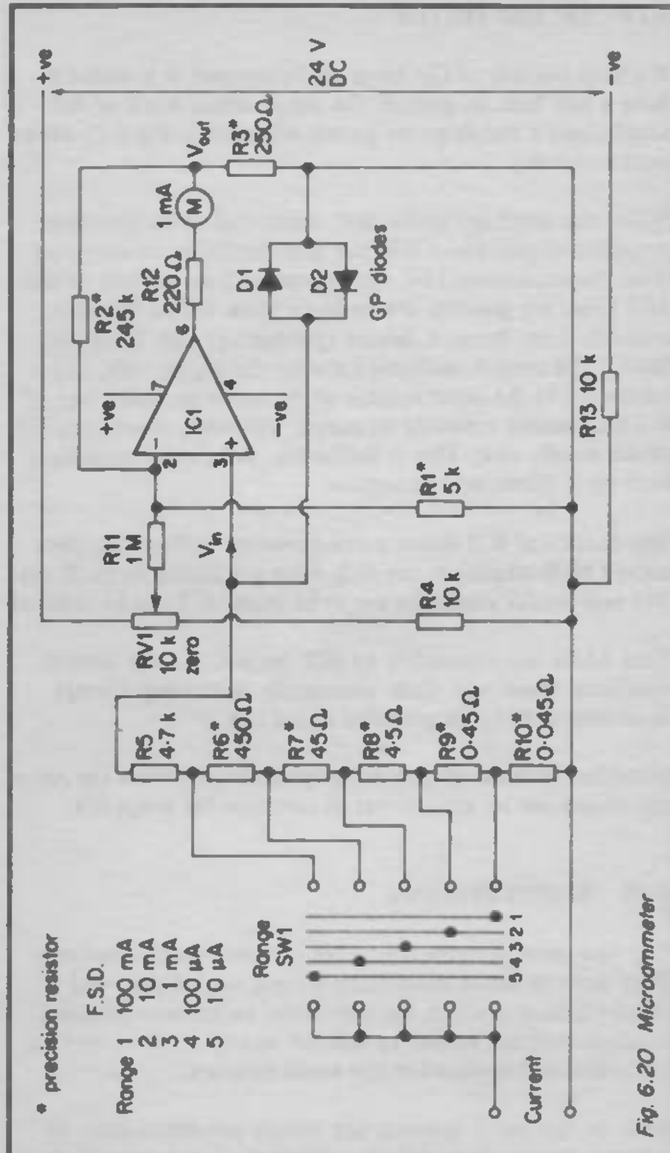


Fig. 6.20 Microammeter

6.14 OP AMP TESTER

If a large number of Op Amps are being used, it is useful to have a test box. In general, Op Amps either work or fail totally, and a simple go/no go test will suffice. Fig.6.21 shows such a circuit.

IC2 is the amplifier under test, connected as an inverting amplifier of gain about 0.8. IC1 is an oscillator, constructed from the ubiquitous 555, with period of 1 sec (details of the 555 timer are given in the author's book IC555 Projects, available from Bernard Babani (publishing) Ltd; Book No. BP44). The output oscillates between the supply rails, and is connected to the input resistor of the amplifier under test. If IC2 is operating correctly its output will swing to within a volt of the supply rails. This is deliberate, since some amplifiers latch up if driven into saturation.

The output of IC2 drives a non-inverting buffer IC3, since some CMOS amplifiers can only drive a milliamp or so. If any 741 and similar amplifiers are to be tested IC2 can be omitted.

Two LEDs are connected to IC2 output. As the output oscillates these will flash alternately indicating correct operation of IC1, the amplifler under test.

Note that because of the pin compatability between Op Amps, the circuit can be used to test all common Op Amps ICs.

6.15 SERVO SYSTEMS

The two servo systems below are position control systems. They have an input comprising a hand turned dial, and a remote indicator which the unit drives to the same position. Position controls similar to this are widely used for remote indication and application like aerial rotators.

Both of the servo systems use simple potentiometers to measure the position of the input unit (called the transmitter,

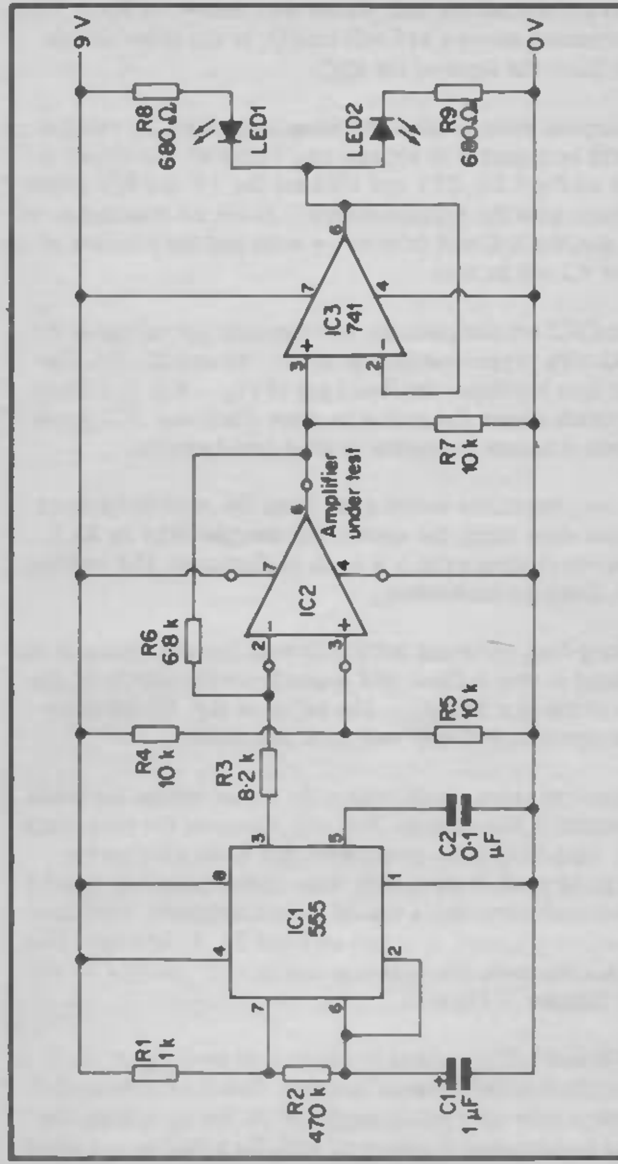


Fig. 6.21 Op amp tester

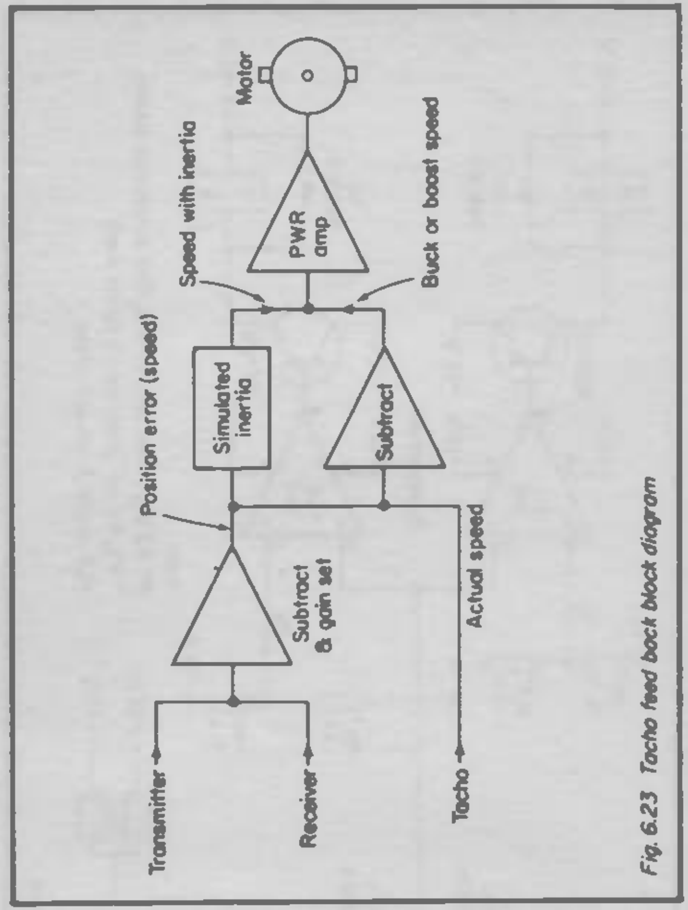


Fig. 6.23 Tacho feed back block diagram

(measured by a tachometer) and a speed error signal produced to buck or boost the actual motor volts. This reduces considerably the tendency of the circuit to oscillate. In the original circuit a slot car motor was used for the main drive motor, and a small motor from a battery toy for the tacho.

RV5 is a zero control used to take out any misalignment, and RV1 the gain control. RV2 calibrates the tacho, and is set such

that zero volts are obtained out of the speed error amplifier with the motor running off load. RV3 sets the tacho gain. RV1 and RV3 should be set by trial and error to give the best response without overshoot.

By comparison between Fig.6.23 and Fig.6.24 the purpose of each IC in the full circuit diagram should be readily available. IC3 and C2 simulate the inertia of a heavy load for demonstration purposes.

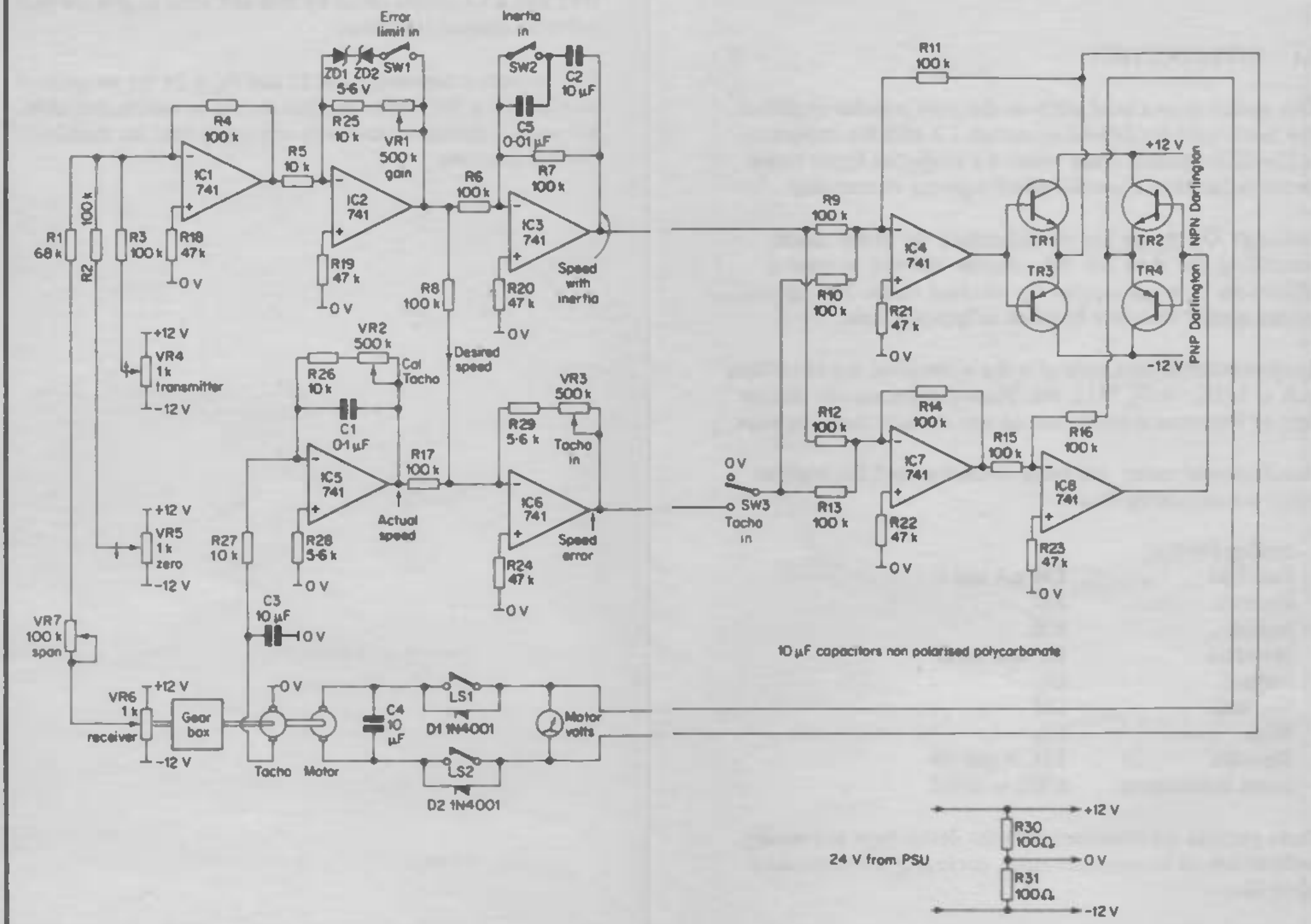
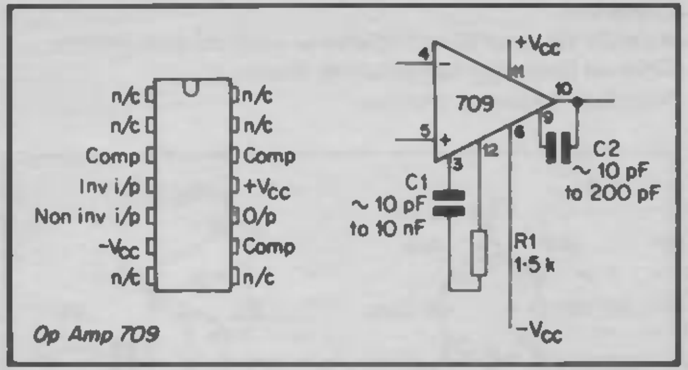


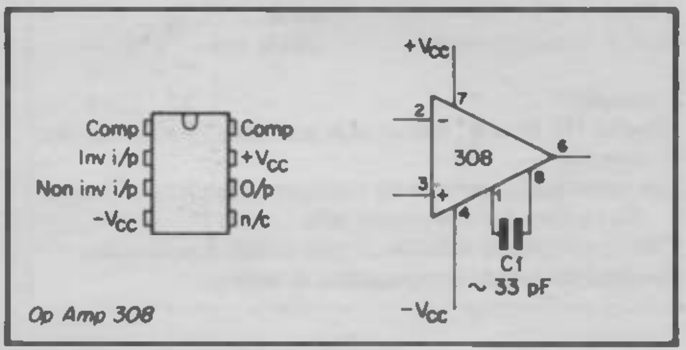
Fig. 6.24 Proportional servo system



Op Amp 709

7.2.5 308

Supply voltage	±2V min, ±18V max
Max differential input voltage	30V
Open loop gain	102dB
Input resistance	40M
Offset voltage	7mV
Offset current	1nA
Bias current	5nA
Slew rate	0.2V/μS
Offset voltage temp coeff	5μV/°C
CMRR	100dB
Useful frequency range	10kHz



Op Amp 308

Comments

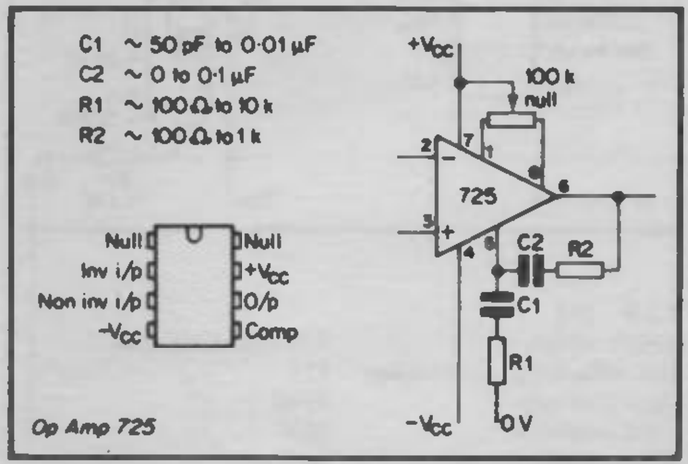
Low drift, instrumentation amplifier;
External frequency compensation as shown.

7.2.6 725

Supply voltage	±4V min, ±20V max
Max differential input voltage	5V
Open loop gain	127dB
Input resistance	1.5M
Offset voltage	2mV
Offset current	1.2nA
Bias current	80nA
Slew rate	0.25V/μS
Offset voltage temp coeff	2μV/°C
CMRR	115dB
Useful frequency range	1MHz

Comments

High performance, low drift amplifier, with correspondingly inflated price;
External frequency compensation as shown.



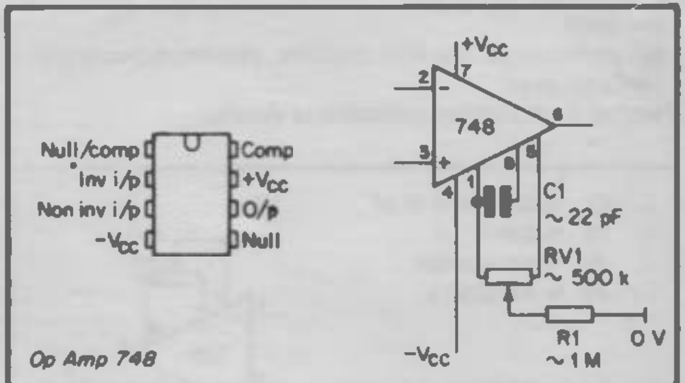
Op Amp 725

7.2.7 748

Supply voltage	±3V min, ±18 max
Max differential input voltage	30V
Open loop gain	106dB
Input resistance	2M
Offset voltage	2mV
Offset current	20nA
Bias current	80nA
Slew rate	0.5V/μS
Offset voltage temp coeff	5μV/°C
CMRR	90dB
Useful frequency range	50KHz

Comments

Similar to the 741 with external frequency components;
Low noise.



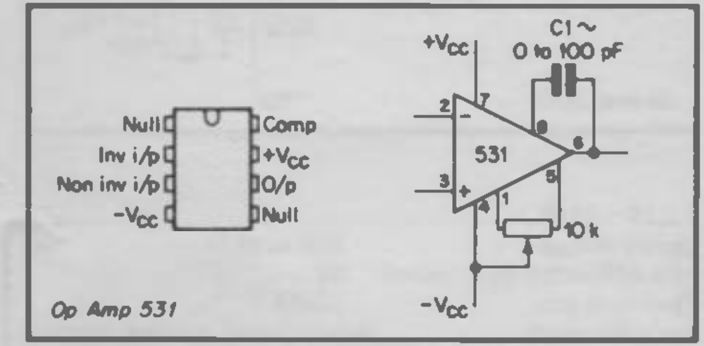
Op Amp 748

7.2.8 531

Supply voltage	±5V min, ±20V max
Max differential input voltage	15V
Open loop gain	96dB
Input resistance	20M
Offset voltage	2mV
Offset current	50nA
Bias current	400nA
Slew rate	35V/μS
Offset voltage temp coeff	10μV/°C
CMRR	100dB
Useful frequency range	> 1MHz

Comments

Very fast slew rate and large frequency range;
External frequency compensation.



Op Amp 531

7.2.9 3130

Supply voltage	±3V min, ±8V max
Max differential input voltage	8V
Open loop gain	110dB
Input resistance	infinite for all practical purposes
Offset voltage	8mV
Offset current	0.5pA
Bias current	5pA
Slew rate	10V/μS
Offset voltage temp coeff	10μV/°C
CMRR	80dB
Useful frequency range	50kHz

Comments

MOSFET amplifier;
External frequency compensation;

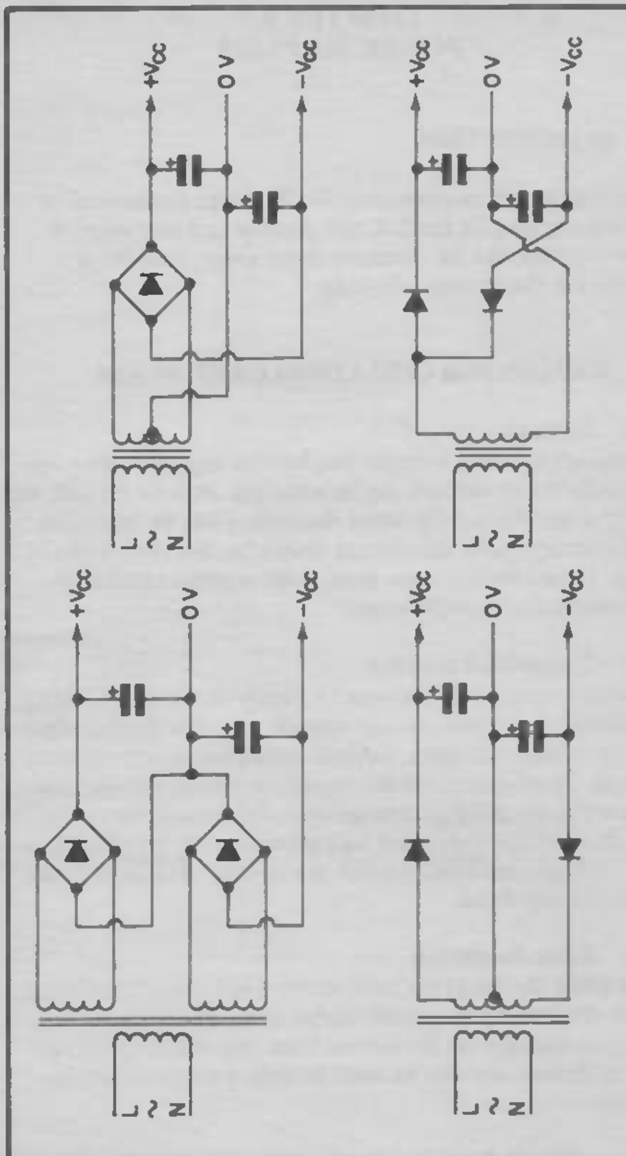


Fig. 8.1 Twin supplies from various transformers

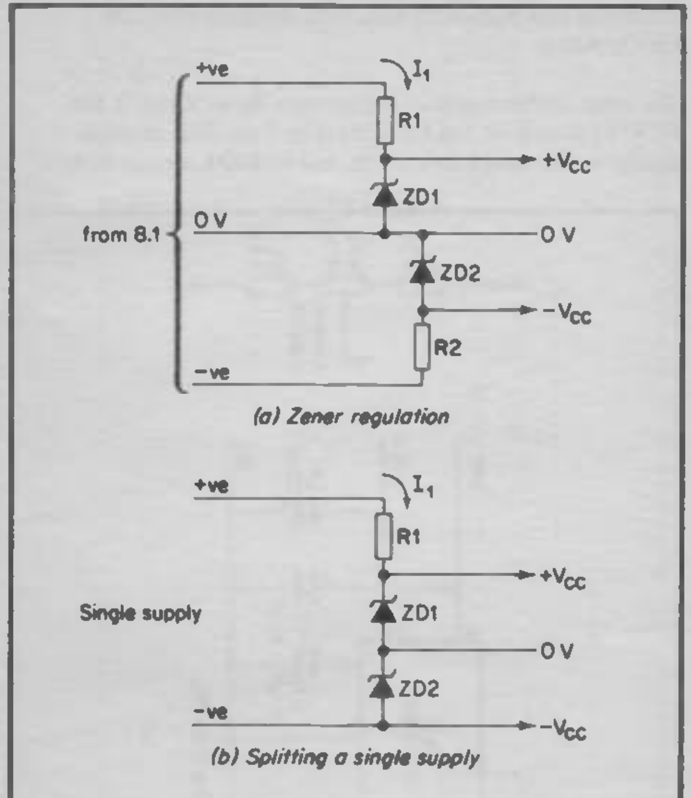


Fig. 8.2 Zener diode supplies

and should be chosen such that I_1 is greater than the maximum load current. In the no load condition, all the current I_1 will flow through the zener, which will dissipate $I_1 \cdot V_Z$ watts. The minimum load thus determines the zener wattage needed.

8.2.4 Integrated Circuit Regulators

The introduction of IC regulators has simplified power supply design to the point where only the most dedicated designer would build one with discrete components. It is not

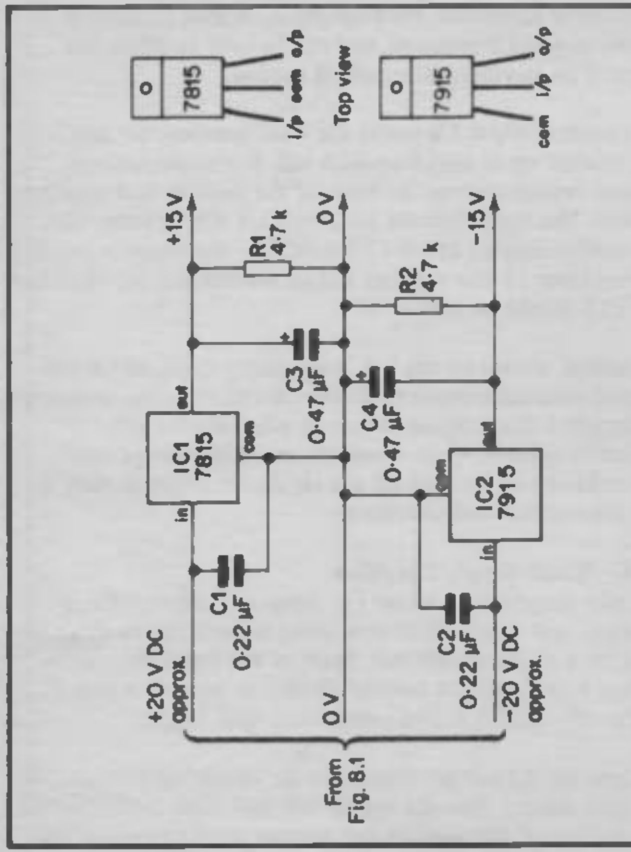


Fig. 8.4 One amp power supply

aries, and Fig.8.5b, whilst technically feasible, is not economically justifiable!

8.2.6 DC to DC Converters

Op Amps are often used with logic, so there is frequently a need for a 5V, +15V, and −15V supply together. This is often best obtained by using a single 5 volt power supply and a DC to DC converter operating off the 5 volt supply to provide the Op Amp rails. Although the DC to DC converter can be home built, encapsulated units are available at a very reason-

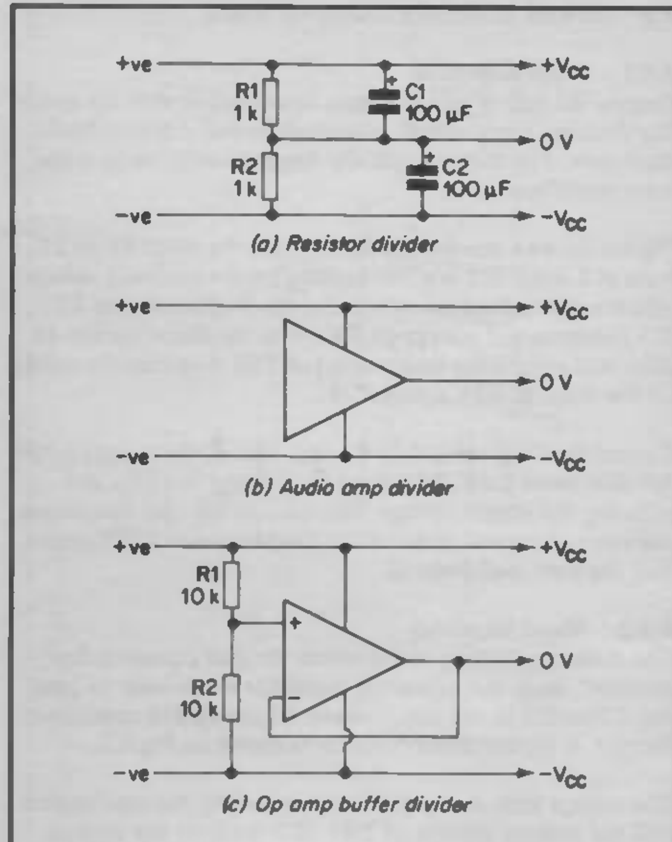


Fig. 8.5 Splitting a single supply

able price (around £5 at the time of writing). The use of these converters with a commercial 5 volt supply is often cheaper than building a three rail power supply.

8.3 POWER SUPPLIES USING OP AMPS

8.3.1 Series Regulator

Despite the ease of power supply construction with the special ICs there are many rugged individualists who prefer to build their own. For these people Op Amps provide ready made error amplifiers.

Fig.8.6 shows a variable supply covering the range 5V to 24 volts at 1 Amp. IC1 is a 741 running on the incoming voltage, which supplies the base current for the Darlington pair TR1. IC1 compares the voltage on RV1 with the Zener voltage on ZD1, and adjusts the base current of TR1 such that the voltage on the slider of RV1 equals ZD1.

Current limit is provided by R5 and TR2. If the voltage across R5 rises above 0.8V, TR2 turns on shorting out ZD1 and reducing the output voltage. The value of R5 thus determines the value of current at which the limit operates. LED1 shows that the limit has operated.

8.3.2 Shunt Regulator

The shunt regulator is useful where the load current is fairly constant, since the regulating transistor needs only to pass the difference in current between minimum and maximum current. A typical shunt regulator is shown on Fig.8.7.

The voltage drop across R2 is determined by the load current and the emitter current of TR1. IC1 controls the emitter current by comparing the voltage on RV1 slider with the Zener voltage on ZD1. The circuit thus compensates for changes in load current and maintains a constant output voltage.

If the load current is fairly constant, the current through TR1 is quite small and low power transistors can be used. Note, however, that the dissipation in TR1 rises with decreasing load, and is at maximum when the load is removed. Shunt regulators should therefore be switched between the regulator and transformer/rectifier if a low power transistor is used for

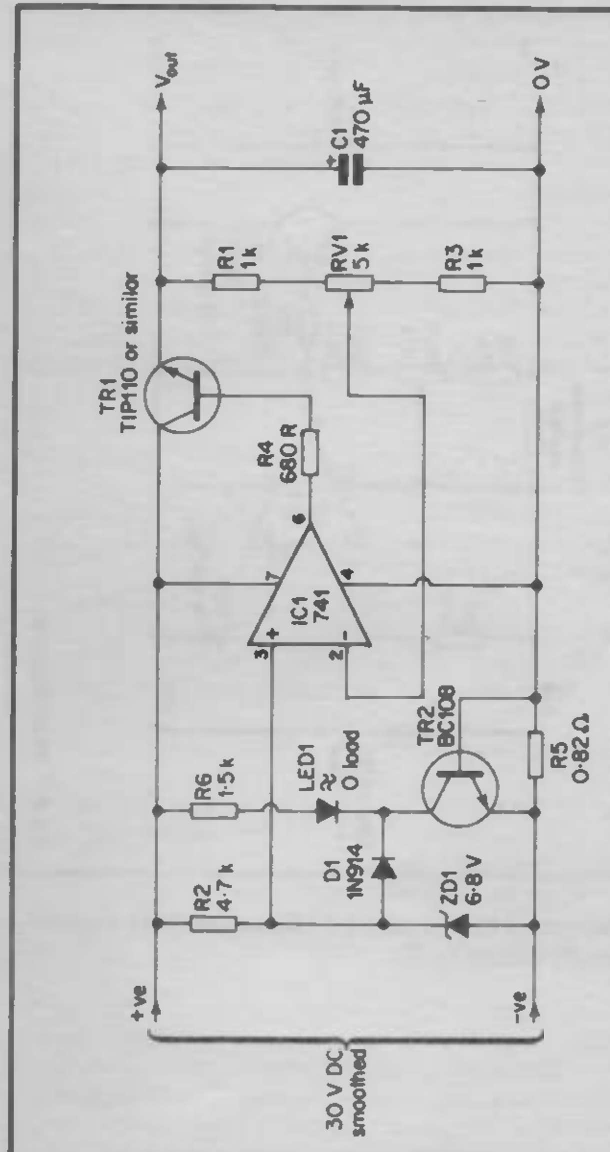


Fig. 8.6 Series regulated supply

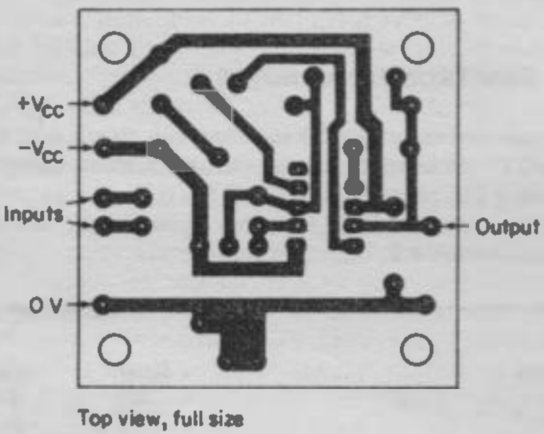


Fig. 9.2 G.P. Op Amp board (Courtesy of R.S. Components)

fessional looking manner.

It is strongly recommended that sockets are used for all ICs. Not only does this simplify fault finding, but most suppliers will NOT replace faulty chips under warranty if they have been soldered in a circuit and removed again.

Op Amps are amongst the most bullet proof ICs, but care should be taken with FET versions. Simple precautions against static should be taken, by resting the PCB on earthed aluminium foil, using an earthed soldering iron and inserting the FET amplifiers only when all other components are in place.

Before turning on a circuit for the first time have a quick visual check for solder splashes, uncut tracks and chips in the wrong way round. We all do it from time to time!

9.2 FAULT FINDING

When a circuit built with ICs fails to work, most people assume that the chip has failed. It should be emphasised that faulty ICs are very, very rare, and dead chips are usually murdered rather than die of natural causes. It follows that replacement of ICs as a first step in fault finding is not recommended, as it could easily lead to the demise of another set of chips.

The first stage of fault finding should always be a careful visual examination, even on a circuit that has worked previously. Points to check are dry joints, shorting components, cracked tracks and the like. On a newly constructed circuit check for solder splashes, uncut tracks and easily made errors like putting a link one hole out or a diode wrong way round. (A common mistake is putting DIL ICs in wrong way round!)

If a visual inspection reveals nothing, its time to get out a multimeter (all electronic engineers should possess a meter, however humble!). First check the supplies for both presence and stability. If these are correct, check the supplies on all the ICs (positive on pin 7, negative on pin 4 on most 8 pin Op Amps).

The next thing to check is the voltage at the inputs and output. These will vary according to the actual circuit, but as a general rule in all Op Amp circuits using feedback, the voltages at the two inputs should be the same. In circuits with high value resistances, the meter may load the circuit, so a meter should be used with a modicum of common sense.

If the voltages do not reveal the fault, now is the time to change ICs. This will be simplified if sockets have been used as advised earlier. If the Op Amp tester of section 6.14 has been built, the removed Op Amp can be tested. If not, another Op Amp can be substituted.

It should be noted that many circuits with adjustments for